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AVIATION ELECTRONICS TECHNICIAN 1&C

NAVY
TRAINING
COURSES
NAVPERS 10318

AVIATION
ELECTRONICS TECHNICIAN
1 & C

Prepared by

BUREAU OF NAVAL PERSONNEL

NAVY TRAINING COURSES

NAVPERS 10318

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U.S.S.D.

THE UNITED STATES NAVY

GUARDIAN OF OUR COUNTRY

The United States Navy is responsible for maintaining control of the sea and is a ready force on watch at home and overseas, capable of strong action to preserve the peace or of instant offensive action to win in war.

It is upon the maintenance of this control that our country's glorious future depends; the United States Navy exists to make it so.

WE SERVE WITH HONOR

Tradition, valor, and victory are the Navy's heritage from the past. To these may be added dedication, discipline, and vigilance as the watchwords of the present and the future.

At home or on distant stations we serve with pride, confident in the respect of our country, our shipmates, and our families.

Our responsibilities sober us; our adversities strengthen us.

Service to God and Country is our special privilege. We serve with honor.

THE FUTURE OF THE NAVY

The Navy will always employ new weapons, new techniques, and greater power to protect and defend the United States on the sea, under the sea, and in the air.

Now and in the future, control of the sea gives the United States her greatest advantage for the maintenance of peace and for victory in war.

Mobility, surprise, dispersal, and offensive power are the keynotes of the new Navy. The roots of the Navy lie in a strong belief in the future, in continued dedication to our tasks, and in reflection on our heritage from the past.

Never have our opportunities and our responsibilities been greater.

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PREFACE

This book is written for Aviation Electronics Technicians of the United States Navy and Naval Reserve who are preparing for advancement in rating to AT1 or ATC. It is one of a series of NAVY TRAINING COURSES designed to give enlisted men pertinent information necessary for the proper performance of the duties of their rate.

The predominant factor in selection of content for this publication has been the *Manual of Qualifications for Advancement in Rating,* NavPers 18068 (Revised), as it relates to AT1 and ATC. Chapters 1 and 12 contain information concerning the administrative and clerical duties of the AT1 and the ATC as set forth in the manual. Chapters 2 through 11 present required technical information that supplements that covered in the following NAVY TRAINING COURSES:

Basic Electricity, NavPers 10086
Basic Electronics, NavPers 10087
Aviation Electronics Technician 3 & 2, NavPers 10317

As one of the NAVY TRAINING COURSES, this book was prepared by the U.S. Navy Training Publications Center, Memphis, Tennessee, and the Navy Training Publications Section of the Bureau of Naval Personnel. Credit also is given to the Aviation Electronics Technician School, Memphis, Tennessee, for preparation of the end-of-chapter questions which review the contents of each chapter of this course.

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NAVY TRAINING COURSES

Basic Electricity, NavPers 10086
Basic Electronics, NavPers 10087
Aviation Electronics Technician 3 & 2, NavPers 10317
Naval Electronics, Part III, NavPers 10810
Physics for Electronics Technicians, NavPers 10095
Essentials of Mathematics for Electronics Technicians, NavPers 10093–A
Advanced Mathematics for Electronics Technicians, NavPers 10094
U. S. Navy Safety Precautions, OpNav 34P1 (ch. 18)
Standard First Aid Training Course, NavPers 10081

USAFI TEXTS

United States Armed Forces Institute (USAFI) courses for additional reading and study are available through your Information and Education Officer.* A partial list of those courses applicable to the AT rating is as follows:

Self-Teaching	Correspondence	Title
	CA 888	*Radio Communication I*
	CA 889	*Radio Communication II*
MA 890		*Principles and Practices of Radio Servicing*
MA 892	CA 892	*Frequency Modulation*
MB 785	CB 785	*Electrical Measuring Instruments*
MB 858	CB 858	*The Slide Rule*
MB 893	CB 893	*Television*

*"Members of the United States Armed Forces Reserve components, when on active duty, are eligible to enroll for USAFI courses, services, and materials if the orders calling them to active duty specify a period of 120 days or more or if they have been on active duty for a period of 120 days or more, regardless of the time specified in the active duty orders."

STUDY GUIDE

Since this training course has been written primarily for the AT General Service Rating, all its contents are pertinent to the GSR. Most of its contents are also pertinent to the qualifications required for the Emergency Service Ratings ATR, ATS, and ATN. The exceptions are that ATS's and ATN's may omit the information on TR and ATR tubes and waveguides in chapter 5, and magnetrons in chapter 6. The ATN may also omit the information on klystrons in chapter 6.

AVIATION
ELECTRONICS TECHNICIAN 1 & C

CHAPTER 1

DUTIES AND RESPONSIBILITIES

As a part of the Navy's training program, this Navy Training Course is written for the purpose of aiding you to prepare for advancement in rating. In order to meet the requirements for advancement you must have gained a wide background of practical experience in the field of aviation electronics theory, and you must also possess a thorough knowledge of electronics from the standpoint of theory. Your past years of naval experience have afforded an opportunity to equip yourself with much of the knowledge and many of the skills that will be of valuable assistance as you progress in your naval work. In addition you should become thoroughly familiar with the contents of the Navy Training Courses given in the preface of this book for these are essentially prerequisites to the publication that you are now going to study.

This training course is meant to be useful for three rates:

1. It is a reference source to refresh the memory of the Chief Aviation Electronics Technician.
2. It is an outline of duties for the Chief Aviation Electronics Technician and for the Aviation Electronics Technician First Class.
3. It is a course of study for the Aviation Electronics Technician 2 who wants to become an Aviation Electronics Technician 1, and for the Aviation Electronics Technician 1 who wants to become a Chief.

MILITARY LEADER AND TECHNICAL SPECIALIST

Your advancement to first class or chief will carry with it many additional duties and responsibilities that have not been required of you to such a high degree in the past. One of the most important is that of training men in lower rates. In order to do this you must possess a firm command of the basic requirements of the rating as well as the personal characteristics that will sell you to the men you are instructing.

You know that as a Navy petty officer you are both a military leader and a technical specialist. In your work, of course, these two phases go together. As a petty officer, you have certain general or military duties; and as a specialist, you have certain professional or technical duties. And you are responsible for attaining proficiency in both phases of a petty officer's work. This book takes up only your specialist or technical (professional) duties. For your military duties, you should study the *General Training Course for Petty Officers,* NavPers 10055. Military requirements are those generally applicable qualifications which all enlisted personnel are expected to demonstrate as a minimum for advancement to specific pay grades.

PRACTICAL FACTORS AND EXAMINATION SUBJECTS

To qualify for a rate you must satisfy the requirements as set forth in the *Manual of Qualifications for Advancement in Rating,* NavPers 18068 (Revised) (quals). These qualifications are printed in the appendix of this course. By becoming familiar with these you will have a better idea of what is required of the different rates of your rating. Of particular importance will be the qualifications that you must possess for advancement to your higher rate.

You must qualify for advancement both with respect to PRACTICAL FACTORS, on which you are tested by performance on the job, and EXAMINATION SUBJECTS, on which you are tested by written examinations. Now, although no book is an adequate substitute for the practice that makes for proficiency in performance, this book, or those cited in it as references, can give you background information that will help you learn more quickly and make you more efficient in the performance of practical work. To the extent that is feasible in this way, then, this book can help you with respect to practical factors.

With respect to examination subjects, this book can give you even more help. What you must do is study these pages and those of the references, and make their contents part of you. Turn to the appendix and read over the qualifications. This book was designed specifically to prepare you to meet those qualifications; you can use them as your guides in planning your study program. This book is based particularly on those qualifications that apply to Aviation Electronics Technician First Class and Chief, but you are responsible also for reviewing the material applicable to lower rates.

In addition, available materials that will enable you to gain further information concerning aviation electronics should be consulted. This material may be obtained from numerous publications, many of which are listed in this chapter under "Sources of Information." Keep up to date electronically by consulting any and all publications, military or otherwise, as often as possible, in order to become cognizant of latest developments.

Talk with others in the field of electronics. Ask them questions; relate to them any unusual electronics finding. Thus, each will profit from the others experiences. Try to learn something new every time you work with your equipment. Analyze your own experience from the practical knowledge you have gained from actually working with electronic equipment.

For general study suggestions on self-study refer to the Navy Training Course, *General Training Course for Petty Officers,* NavPers 10055, chapter 8.

YOU AS AN AT1

Advancing from AT2 to AT1 means much more than an increase in pay and the acquiring of the privileges of a first class petty officer. This step up the Navy ladder of success means that you must increase your knowledge and skills as they relate to your field. You must assume the responsibility of training men as they advance in the field of electronics. Also, it means that much of your time will be devoted to work of a supervisory nature, and in doing this you will be required to assume duties that are somewhat new. You will be required to make decisions, for in many situations you will be in charge of the electronics department of an activity or squadron.

YOU AS AN ATC

Advancement to chief petty officer is far from a routine promotion. This rate carries with it increased authority and responsibility, and requires the exercise of more leadership and initiative than previous rates.

A list of the more important duties, with a brief explanation, is presented as a helpful reminder to aid you in gaining an understanding of the extent of your responsibilities as a Chief Aviation Electronics Technician.

Supervision

An ATC must be able to take over management of the electronics shop. It will be your job to assign work and see that it is properly executed. You must check to insure that the work is being done according to predetermined standards and that no inferior work is permitted.

You may serve as division chief and be responsible for both the electronics and the electrical shops. Take steps to insure that all hands have been properly instructed regarding safety precautions to be employed in electronics work.

On-the-Job Training

You will also be responsible for seeing that the lower-rated petty officers and strikers in your shop receive the proper training. This requires thought and planning.

For example, there may be times when the men in your shop will be required to work 12 or 16 hours a day to complete a rush job. At other times you have nothing but routine work for several days. That's when you'll find time to train your men.

When the equipment in the shop is not required for the regular work, assign some practice jobs to your strikers and lower-rated men. Give them a chance to learn new skills.

Pay particular attention to the training of men who are weak in one or two phases of the work. If a man lacks skill in a certain type of work, give him a chance to practice it alongside a skilled worker.

Do not use hit-or-miss methods of teaching. Give the man a definite job to do and check him occasionally on the job. Give him all the help you can. Try to plan his training so that he can acquire the most knowledge and skill in the least amount of time.

When one of your better men is working on an unusually difficult job, have some of the other men watch the operation. This procedure is invaluable from the standpoint of training, and it also serves as a well-deserved compliment to the man who is doing the work.

Personnel Problems

As a chief, you will frequently become involved in personnel problems, such as recommending men for promotion, and settling differences and grievances between them. You must handle such matters with discretion and fairness. Your job will put you in the role of a liaison man, for you will have to deal with equals and superiors as well as subordinates. Your attitude toward this chain of command is important since the whole structure of naval service is built upon ascending and descending responsibilities. Each officer and man in the Navy is dependent upon others for the smooth performance of his tasks.

The morale of men under you will be an index to the efficiency of your division. Attempt to detect small problems before they grow into large ones. Counsel with your men, give them advice and assistance whenever possible.

Records and Reports

An important part of your duties is paperwork. As a leading petty officer, you are required to keep many records and submit many reports. A more detailed description of these duties will be presented later in this chapter.

SQUADRON OR ACTIVITY ORGANIZATION

DIVISIONAL COORDINATION.—Working together is important to an organization and it is especially meaningful to a squadron. With the dependence of the divisions on each other there must be cooperation if the job is to be done. You must be able to work with men of other divisions, and must be broadminded enough to understand their problems; for on numerous occasions you may have to change work plans in order to allow others to complete necessary tasks. If you follow this plan with a smile and a sense of responsibility, you will receive dividends for you will certainly have to ask favors of others from time to time.

ELECTRONICS DIVISION OFFICER.—This officer has the responsibility of supervising the operation and upkeep of all aircraft electronic equipment attached to his activity. He also supervises the operation and upkeep of the facilities for servicing and testing this equipment. He is your immediate supervisor and as such must rely upon you to keep him posted as to the overall functioning of the electronics division. To a degree, his responsibilities are yours; and without a doubt, your responsibilities are his. Learn to work with your officers, understand their position, and bring to them problems that you are unable to solve. Through such action you will do much to gain a higher mark for the electronics division.

Maintenance Department

The maintenance department is organized according to instructions issued by the Bureau of Aeronautics,

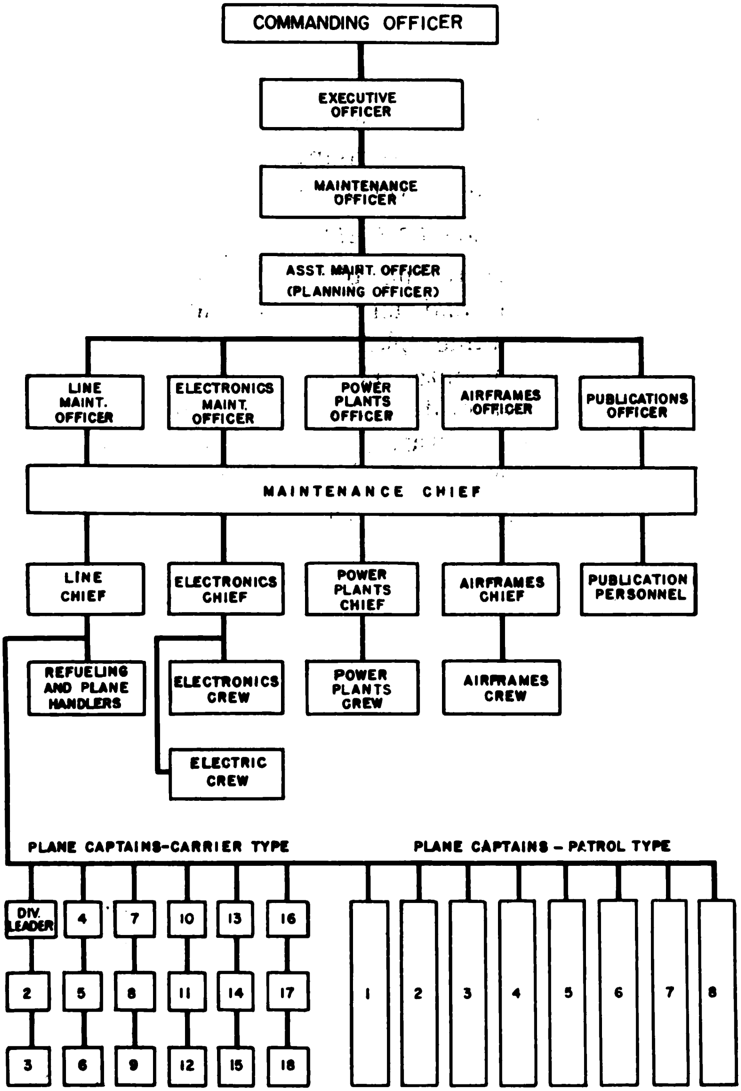


Figure 1–1.—Organization chart, typical squadron or activity.

Naval Air Station Commander, or the specific Naval Fleet Air Force Commander to which the squadron or activity is attached. (See fig. 1–1.) Maintenance activity is a term with broad meaning for it applies to available shop equipment, shop space, or hangar space used for aircraft overhaul, modification, routine upkeep or servicing, and the personnel required for such operation.

The far-reaching duties of this department may be more fully understood by studying the following responsibilities as they relate to various squadrons:

SUPPORTED SQUADRON RESPONSIBILITIES:

1. Squadrons are responsible for the material condition of all aircraft and aeronautical material in their custody, and all routine upkeep and servicing within the capabilities of the maintenance facilities available. They should insure incorporation of required service changes and maintain their aircraft in the highest possible material readiness for war.

2. Squadrons should perform the following inspections and correct or cause to be corrected all discrepancies on assigned aircraft:

Preflight.
Postflight.
Preembarkation.
Weekly material inspection.
Periodic maintenance inspections.
Transferring and receiving, including preservation and depreservation as directed.

3. The squadron commanding officer should insure the optimum level of maintenance training through the widest possible diversification of duties and delegation of responsibility with adequate tutelage and technical supervision.

4. Squadrons should maintain a technical library of all pertinent instructions for the operation, routine upkeep, and servicing of the type aircraft assigned.

FASRON RESPONSIBILITIES:

1. The FASRon should provide the supported squadrons with spaces and furnishings and authorized facilities and equipment required for maintenance and servicing.

2. The FASRon should act through the fleet air commander as the agent for the squadron in securing overhaul and repair services for squadron work beyond the capability of the squadron and FASRon.

3. The FASRon is responsible for the material condition of its shop spaces and equipment and provides instruction and supervision where necessary for squadron personnel utilizing these spaces and equipment.

4. The FASRon should maintain a technical library of all competent instructions pertinent to the operation, routine upkeep, and servicing of all aircraft types assigned to fleet air tactical units they are designed to support.

5. The FASRon is responsible for assisting the squadron with its maintenance during periods of "peak loads" resulting from accelerated programs and in special work requiring highly skilled mechanics or technicians not available in the squadron.

6. The FASRon is responsible for the material condition of all assigned and pool aircraft and aeronautical material in its custody, and the responsibilities in connection therewith are the same as those set forth for squadrons.

SELF-SUPPORTING SQUADRONS:

The maintenance mission of self-supporting squadrons combines the responsibilities listed for FASRon and supported squadrons. (Refer to the above listings.)

Overall efficiency of a squadron or activity may be measured by the coordination of the various divisions. The maintenance chief is responsible for the readiness of the aircraft; therefore he must work with all division leading petty officers and they in turn must cooperate.

Any change in aircraft status must be reported to the maintenance chief in order for him to do his job satisfactorily. As an example, if you wish to remove a piece of equipment from an aircraft for checking or bench repairs, it is important to notify the maintenance chief. The plane may have to be placed in a DOWN status; if so, the chief may then notify operations. Also, any other shop that may desire to work on the plane may be notified.

ELECTRONICS DIVISION.—The electronics division is charged with the responsibility for the operation and upkeep of facilities for servicing, testing, and proper functioning of aircraft electronic equipment.

These duties include the detailed supervision of the care, adjustment, maintenance, repair, installation, and security of all airborne equipment or devices which depend upon electric current for their operation in aircraft. Some of these equipments are: communications, homing, radar identification, radio and radar countermeasures, search radars, radar bombing attachments, blind landing aids, generators, motors, voltage regulators, reverse current relays, batteries, lighting, electric meters, gyro stabilized compasses, automatic pilots, instruments, and electrical wiring.

The electronics division may be under the direction of one or more officers who come directly under the maintenance officer. It is important that the Electronics Officer be informed of any action, other than routine, pertaining to his division. Without such information divisional coordination will become almost impossible and the squadron or activity will not measure up to its operational commitments.

Types of work performed by an electronics division are determined by the organizational classification as set forth by the Bureau of Aeronautics. (NOTE: For detailed information read BuAer Instruction, NavAer 00.62.) All land-based maintenance activities are classified under six major classes. The degree of overhaul or repair work which may be performed by the aeronautical maintenance activity is limited by the classification of the maintenance activity. These classifications are as follows:

Type	*Alphabetical Classification*	*Level*
Overhaul	A	Overhaul and Repair
Overhaul	B	Overhaul and Repair
Maintenance	C	Shop/Repair
Maintenance	D	Shop
Maintenance	E	Hangar
Maintenance	F	Line

Most of the land-based electronics work performed by Navy personnel in the AT rating will be under the C and D classifications. The degree of maintenance work to be performed on electronic equipment is the same for class C and class D activities.

Electronics aircraft maintenance work that may be performed at class C and class D activities and on board aircraft carriers, tenders, battleships and cruisers is as follows:

1. Perform routine maintenance check of all installed electronic equipment.
2. Perform flight and ground tests of electronic equipment.
3. Repair or replace defective wiring, cabling, antennas, radomes, etc.
4. Calibrate equipment while installed in aircraft if possible, otherwise, remove to shop for calibration.
5. Test and replace as necessary, brushes, tubes, resistors, capacitors, lamps, and fuses, or such other parts normally provided on the allowance lists.
6. Perform routine maintenance on shop tools and test equipment.
7. Calibrate electronic test equipment using "WWV" standard frequency transmissions.

A detailed explanation of established general policy and delegated responsibilities for electronics maintenance may be found in BuAer Instruction 10550.2A.

TECHNICAL DUTIES OF AVIATION ELECTRONICS TECHNICIANS

One of the more important tasks you will have as a first class or chief is to see that the men under you follow through on all matters pertaining to the division. When there is an electronics discrepancy, see that the work is expedited; and if a unit must be replaced, be sure to repair the defective unit as soon as possible. If a part is needed for repair, and if it is not locally available, order it immediately to insure adequate equipage availability. A unit that is awaiting repair should be properly identified. Tag the unit, showing the nature of repair needed and action taken, such as part or parts ordered, date, and person responsible for such action.

For adequate use, test equipment must be in proper working order. In many shops this equipment is allowed to become inoperative because of misuse and nonuse for long periods of time. Due to the nature of much of the equipment it is important that periodic checks be made to insure proper operation. This is especially true of equipment that makes use of batteries. When test equipment is found inoperative be sure to tag it and show the cause of trouble. This will prevent others from attempting to use the equipment and will save time and effort.

REPAIR OF AIRCRAFT ELECTRONIC EQUIPMENT.—Hand tools will pose a problem for the electronic shop in most cases for it seems that there are never enough to fill the needs. This is usually a result of improper organization for the tools are issued according to the needs of the activity and these needs are established by competent authority. Improper storage, disregard for responsibility, improper use, and carelessness are the major causes for "hand-tool headaches." Petty officers in charge may easily increase divisional efficiency by taking time to develop a workable method of proper care and use of hand tools.

TEST EQUIPMENT REPAIR.—All activities other than designated overhaul points shall make only minor repairs to units of defective electronic and electrical (classes 16 and 17) test equipment. The term "minor repairs" is defined as the replacement of such parts which will not affect the circuitry or calibration of the test equipment. These parts are as follows:

1. Electron tubes, crystals, fuses, batteries, and component parts which are commonly available and listed in the NavAer 00–35QR–4 allowance list.
2. Peculiar parts listed in the NavAer 00–35QR–5 allowance lists.

A brief description of any such minor repair made should be included in the remarks section of the applicable Electronics Failure Report.

Most test equipments have a high degree of accuracy, are inherently delicate in nature, and require extreme care and caution in handling in order that their calibration accuracies remain undisturbed. Signal generators and frequency meters should be periodically calibrated in accordance with current directions.

AUXILIARY POWER EQUIPMENT.—The leading petty officer should take action to assure that men assigned to his activity are properly indoctrinated in the use, maintenance, and stowage of all electronic auxiliary power equipment. See that all safety measures are observed when using this equipment, especially when applying external power to an aircraft.

Generally speaking, maintenance of this equipment is not the responsibility of the AT. He does have serious obligations for he will most likely be the one who detects trouble; he also must use the equipment in the proper manner. The AT must report equipment failure to the division responsible for maintenance and upkeep. The procedure that he follows should insure that necessary steps are taken to protect the equipment and to facilitate rapid repair. He should promptly tag the unit and then inform the responsible division, being sure to give all pertinent information dealing with the discrepancy. By so doing, the following may well be accomplished: First, prevent another from wasting valuable time in trying to use the defective unit; and second, cause the unit to be placed in operation without unnecessary delay.

Without a doubt, auxiliary power equipment failure has caused the Aviation Electronics Technician almost as many headaches as any other phase of his work. Many planes have been grounded due to inferior auxiliary power equipment. Much divisional misunderstanding has developed due to inoperative and misused equipment. Make provisions to insure against such incidents and you will have a better division.

ELECTRONIC MATERIAL CHANGES.—Electronic Material Changes are issued by the Bureau of Aeronautics to promulgate technical instructions necessary for the addition, removal, or replacement of a part, modification to circuitry, and other necessary changes. Modification of electronic material will normally be limited to modification of parts, subassemblies, assemblies, components or complete equipments, and interconnecting wiring. Instructions for the SUBSTITUTION of components of an electronic system may be issued by an Electronic Material Change, provided the following requirements are met:

1. The change must have negligible effect on weight and balance of the aircraft.
2. The aircraft structure and/or nonrelated equipment(s) must not be affected.

NOTE: SUBSTITUTION is defined as removal of equipment or component and its replacement with a similar equipment or component to perform the same or additional functions.

Modifications which fail to satisfy the above requirements will be handled in accordance with aircraft service change procedures.

Electronic Material Changes should be performed in accordance with the instructions contained in the directive authorizing the change.

The importance in performing these changes promptly and with the utmost of care cannot be overemphasized. They keep your equipment up to date, interchangeable with replacement units, and prevent possible failures. After the change has been completed, check it over for good connections, security of components, and give it a complete functional test.

It will be advisable to refer to the Electronic Material Changes (EMC's) at various times. For example, some of the changes are not to be accomplished until the part or parts have failed. In this case, when a part fails it is advisable to check for any EMC's that may apply to the failure. There may be times when you will find that a circuit will differ from the schematic diagram, and by checking the EMC file it may be found that this difference is due to a change. The *Handbook of Service Instructions,* which contains the schematic, in this case, has not been corrected to comply with the change which accounts for the difference.

The changes that apply to the equipment in your activity may be found in the *Naval Aeronautic Publications Index,* NavAer 00–500. (NOTE: Electronic Material Changes are issued in lieu of Technical Orders.)

ELECTRONIC MATERIAL BULLETINS.—Electronic Material Bulletins are issued by the Bureau of Aeronautics to promulgate technical instructions on operation, inspection, alinement, maintenance procedures, etc. These bulletins are issued in lieu of Technical Notes, and a listing of them may be found in the *Naval Aeronautic Publications Index,* NavAer 00–500.

NEW ELECTRONIC INSTALLATIONS.—Due to the rapid development of electronic equipment you will frequently be called upon to modify old and install new. When installing new gear, the installation instructions contained in the *Handbook of Service Instructions* for the equipment being installed should be consulted for information peculiar to that equipment.

In evaluating an installation it will be necessary to determine that the equipment performs properly. This is done by checking power output, sensitivity, the functioning of all indicators, proper meter readings, and other checks as applicable. There should be a complete check to determine if the equipment interferes with other equipments, or if the other equipments interfere with it. If there is interaction, it may be the result of case radiation, poor bonding, coupling between interconnection cables (similar to crosstalk), or coupling between antennas. Another trouble may be low antenna output due to excessive antenna cable length. If possible, antenna patterns should be made on all new antenna installations to check for possible dead spots (bearings with poor radiation). This can be done by orbiting a plane around a fixed point which is a known distance from the station taking readings to determine the pattern.

When you are required to make an aircraft electronics installation, all of the preceding information on evaluation should be taken into consideration, as well as the following:

1. Exercise great care and supervision in the fabrication of all cables, especially the connections in plugs.
2. Keep all cable lengths as short as possible, especially antenna coaxial cables. Also avoid sharp bends. (There generally are some installation specifications as to maximum permissible cable lengths in the *Handbook of Service Instructions.*)
3. Prevent possible cable chafing by proper lacing and cable straps.
4. Provide proper power circuit protection.
5. Provide for careful placement of individual units. For example, keep high power pulsed transmitters away from receivers, especially those that operate in the same frequency range, or those of high sensitivity.
6. Maintain proper clearance (generally one inch)

between each new unit and adjacent equipment, and structural members.

7. Provide accessibility of units for operation and maintenance.
8. Units should be so located that free air circulation is provided, especially in the vicinity of air intakes to blower motors and exhaust louvers.
9. Surfaces to be bonded must be clean and scraped of corrosion and paint to insure proper electrical ground.

SAFETY

SHOP AND HANGAR AREA.—Supervisor personnel have the following duties pertaining to safety:

1. They shall enforce safety regulations.
2. They shall instruct and drill their men in safe practices, and shall provide and enforce the use of personal protective equipment.
3. They shall carry out recommendations submitted by proper authority for the correction of unsafe conditions and work methods, and shall caution operational personnel regarding hazards beyond their control.
4. They shall analyze accidents within their jurisdiction and apply or recommend corrective action to prevent recurrence.
5. They shall seek advice and assistance from the Safety Officer in their administration of the accident prevention program.

The subject of safety is covered briefly in the Navy Training Course, *Aviation Electronics Technician 3 and 2,* and completely in the publication *U. S. Navy Safety Precautions,* OpNav 34P1.

The following precautions serve to emphasize the more hazardous areas of electrical work:

1. PERSONNEL. Only authorized personnel shall be permitted to install and maintain electrical equipment.

Personnel working with electrical circuits shall avoid wearing loose clothing, rings, watches, and other jewelry.

2. SAFETY DEVICES. Switch panels and power distribution boxes shall be protected by enclosures. Rubber mats or insulated decks shall be provided at operating locations.

3. CONTROL SWITCHES. Personnel shall be certain that switches are off before plugging in external power. Before a switch is thrown it must be ascertained that personnel are clear of the equipment.

4. HIGH VOLTAGES. When working near exposed high voltage do not stand on metal or wet decks, and do not touch the grounded frame of the equipment. Keep yourself above ground potential and you will lessen your possibility of shock. A person should never work alone around electrical equipment.

5. ELECTRONIC POWER SUPPLY CIRCUITS. All power supply filter capacitors should be grounded before work is done on electronic equipment. Grounded shorting prods should be permanently attached to work benches where radar equipment is regularly repaired.

6. DISPOSITION OF RADIOACTIVE TUBES. Useless unbroken tubes containing radioactive material, such as spark gap, TR, glow lamp, or cold cathode-ray tubes, should be treated as any other radioactive waste material. They should be sunk intact at sea. In shore installations it is best to collect the tubes in special containers which should be weighted, sealed, and shipped out to be sunk at a convenient time. A plot of land may be set aside near the shore station to be used as a burial ground. However, the former method of sinking is recommended. If a burial plot is used, it should be adequately posted and supervised.

Any equipment or tools used in crushing tubes or handling radioactive junk should be thoroughly cleaned before using for other purposes, or, if practicable, such equipment and tools should also be buried or sunk at sea. It should be borne in mind that any buried material may at some later date be exposed by land excavation and cause radioactive poisoning exposures.

SAFETY IN AND AROUND AIRCRAFT.—Most of the safety precautions for the shop will also apply to the work on aircraft. The following precautions are generally peculiar only to aircraft:

1. HIGH INTENSITY RADAR. Care must be taken within a 75-foot radius of high intensity radar to prevent ignition of certain substances such as volatile gases.

2. FLAMMABLE FUMES. A check for presence of flammable fumes shall be made before activating electrical or electronic circuits, or keying transmitters.

3. RESPONSIBILITY WHILE IN FLIGHT. Before making any repairs to electronic equipment, other than routine, receive permission to do so from the pilot. Also, obtain permission from the pilot to energize radar and radio transmitters. He is responsible for the safety and operation of the aircraft and as such must be kept informed of any conditions other than normal. Whenever in doubt consult the pilot!

4. FUELING. While an aircraft is being fueled, no electrical equipment will be energized and no electrical apparatus supplied by outside power (electrical cords, droplights, floodlights, etc.) will be in or near the plane.

SOURCES OF INFORMATION

I know that I have read it. It gave me the information that I need. I'll remember the title of the publication. Do you know about so and so dealing with so and so? How can you expect a person to remember everything? More is written about electronics than any other subject, can't expect to remember all of it.

The above statements are probably typical of the many thoughts that cross your mind as you have attempted to recall some bit of information that you needed. You will not be able to retain by memory all of the information dealing with electronics that you need to use from time to time. The following sources of information, with brief descriptive statements, should prove helpful and are given as a ready reference. They will also enable you to become more up to date from the standpoint of your rate.

A suggested method whereby general interest materials, such as the manuals on connectors, safety precautions, noise elimination, etc., may be readily available is to place them in suitable binders and keep them in the shop.

DIGEST—U. S. NAVAL AVIATION ELECTRONICS.—The Naval Aviation Electronics Service Unit each month prepares and publishes the magazine *Digest of U. S. Naval Aviation Electronics*. It gives information on electrical and electronic equipment relative to operation, maintenance, installation, and supply. This is an excellent source of information dealing with new equipment and new ideas on the older equipment. It gives the latest procedures for speeding the testing, calibrating, and alinement of equipment now in use as these methods are developed by the Electronics Service Unit and other activities. In general, it is as the name suggests, a "Digest" of all the latest information pertaining to your work in electronics. It should be on the "must read" list for all hands and always available for ready reference.

ELECTRONIC MATERIAL CHANGES AND BULLETINS.—These are excellent sources of information and you should be familiar with the ones that relate to your equipment.

NAVAER ALLOWANCE LISTS.—Allowance lists are a good source of stock numbers for they are kept up to date by frequent revisions. Time spent becoming familiar with the Section R Allowance Lists is well worthwhile for they are a good source of information. For example, the section on electron tubes, (NavAer 00–35–QR–7), contains charts that show the quantities of each type of tube that each equipment uses.

PUBLICATIONS ALLOWANCE (BuPers Instruction 5601.1).—This instruction sets forth publications allowance for individual ship classes, aircraft squadrons, and staffs. The publications listed are divided into three categories, as follows: Category I—Required; Category II—Recommended; and Category III—Issued upon request.

Directions for procurement of these publications are contained in the instruction. NavPers numbers and titles of all publications are also given.

ILLUSTRATED PARTS BREAKDOWNS.—These breakdowns contain illustrated group assemblies, generally exploded views, together with parts lists for each group. They also contain numerical parts and reference symbol indexes and furnish one of the best means of identifying parts.

NAVAL AERONAUTIC PUBLICATIONS INDEX (NavAer 00–500).—The *NavAer Publications Index* contains a consolidated listing of all publications of interest to aeronautical activities. It may be used as a check list in determining publications requirements and as a guide in ordering publications. Complete ordering information is included for all publications listed.

In order to determine the quantities and types of publications and forms that each naval aviation activity is allotted refer to NavAer Section K Allowance List, NavAer 00–35QK–1.

NOMENCLATURE ASSIGNED TO NAVAL ELECTRONIC EQUIPMENT, NavShips 900,123(A) and supplements, or CO–NavAer 16–45–511.—This book contains descriptive data on all equipments to which Navy model letters and joint nomenclature (AN) model letters have been assigned. As such, it provides a convenient reference by which equipment can be identified if the model letters are known. It contains only the equipment developed by the Navy. This publication also contains a list of contractor's designating symbols as well as a summary of the Joint Nomenclature System ("AN" System) for communication-electronic equipment.

U. S. NAVY SAFETY PRECAUTIONS, OpNav 34P1.—This publication is a rather complete text on the subject of naval safety and is the result of compiling all directives and publications previously issued on the subject. All supervisory personnel should become familiar with the chapters dealing with aviation and the chapter on electricity and electronics.

INSTRUCTIONS AND NOTICES.—A system of dissemination of information by the Navy is important and takes many forms. The above heading is one of the most frequently used methods and is prescribed for the issuance of directive-type releases including those which prescribe policy, organization, methods or procedures, and those which contain information. This program is called the Navy Directives System and provides a uniform plan for issuing and maintaining directives. Two types of releases are authorized under the plan: 1. Instructions, and 2. Notices. The Aircraft Circular Letters, so very much used in the past, are now incorporated into this system. The manner of numbering is described in *Aviation Electronics Technician 3 and 2,* NavPers 10317. The Navy Directives System is thoroughly explained in SECNAV INSTRUCTION 5215.1. This instruction also includes a table of Subject Classification Numbers which may be used as an index for finding information about a particular subject. Another source of information dealing with Instructions and Notices is the Consolidated Subject Index which is issued periodically by the Navy Department, Administrative Office, under Notice No. AO 5215.

CROSS INDEX OF ELECTRON TUBES.—This index, which is a supply publication, lists all electron tubes used by the Navy and gives the following information about them: The stock number for the tube, the type or types of tubes it replaces, and the type or types of tubes that are interchangeable with it.

AIRCRAFT MAINTENANCE HANDBOOKS.—This type of publication is intended to provide information which will aid in the maintenance of aircraft. Information is presented as to location, function, operation, removal, installation, testing, adjusting, and troubleshooting of components. Maintenance methods recommended are concerned with procedures which can be accomplished by operating units.

Before attempting any new task on an aircraft, such as removing a radome, the *Handbook of Maintenance Instructions* for that particular aircraft should be consulted. By proper use of these handbooks possible damage will be prevented and much time may be saved. These handbooks are so arranged that the sections pertaining to electronics may be removed and kept available in the electronics shop. The electronics sections are as follows: Section VII, Electrical Systems; Section VIII, Radio and Radar; and Section X, Wiring Data.

HANDBOOK OF OPERATION AND SERVICE INSTRUCTIONS ON AN ELECTRICAL CONNECTORS, AN 03–5–90.—This handbook covers the description, selection, preparation, installation, and maintenance instructions for the commonly used AN connectors. This handbook may serve as a source of information for technicians and also as a training guide for men inexperienced in the construction of cables using AN connectors.

MANUALS, BUAER AND BUSHIPS.—These manuals are published for the purpose of disseminating information to naval personnel. The information is presented in chapter form with each chapter dealing with a particular topic. Many times these chapters seem to be a source of impossible-to-find information. They are kept up to date by revision, and new chapters are published as new needs arise.

Some of the chapters are of particular importance to the AT1 and ATC. They are listed here for your benefit as a source of readily available information:

1. *BuAer Manual,* Chapter 10, Aircraft Electronic Equipment.
2. *BuShips Manual,*
   Chapter 60, Electric Plant, General

Chapter 61, Electric Generators and Voltage Regulators
Chapter 62, (Section 2) Portable Storage Batteries and Dry Batteries
Chapter 67, Electronics
Chapter 69, Electrical Measuring and Test Equipment

FLEET REGULATIONS, AIR FORCE.—These regulations are a source of information dealing with topics such as organization, duties, training in safety, personnel, aircraft material maintenance, supply and communications, etc. They establish specific duties and responsibilities for a particular organization or activity.

MILITARY SPECIFICATIONS AND STANDARDS, LIST OF.—This is a complete listing of the military specifications and standards used by the Bureau of Aeronautics. It is issued by the Department of the Navy, Bureau of Aeronautics, Washington 25, D. C., and contains information relative to the distribution of and requests for specifications. These specifications may be used as an aid in the performance of a particular installation or test, and they also provide an excellent source of information of a general nature.

An example of information that may be obtained from such specifications as listed in this publication is as follows: Military Specification, MIL–W–5088, *Wiring, Aircraft, Installation of.* This specification covers the installation of electric and electronic equipment in aircraft. It also includes a listing of many other specifications, standards, drawings, and publications, related to this subject. This specification is of particular interest to the AT and should find wide use if made available to shop personnel.

HANDBOOK—INSTALLATION AND MAINTENANCE INSTRUCTIONS, NavAer 16–1–518.—The title of this publication is *Anti-Precipitation Static System.* This handbook explains what must be included in the complete anti-precipitation static system for an aircraft and describes special components which are used, including proper procedures for installation and maintenance. It also includes a table of replaceable parts and instruction sheets to be used in the assembly of antenna fittings.

INDEX—TRANSMISSION LINES AND FITTINGS, NavShips 900–102B (ASESA 49–29A).—This publication contains charts and tables which aid in the selection of proper cables and fittings to be used in constructing coaxial lines.

AIRCRAFT SERVICE CHANGES.—These changes are authorized and published by BuAer for incorporation in aircraft after delivery and acceptance by the Navy. They are for the purpose of improving the flight and safety characteristics of an aircraft.

These changes contain the following information:

The type plane to undergo a change.

The title of the change.

Aids in making the change, such as, photographs, drawings, sketches, and contract bulletins.

Materials to be used in making the change and directions for procurement of such materials.

Changes in weight and balance.

These changes are classified into the following three categories—IMMEDIATE ACTION, URGENT ACTION, and ROUTINE ACTION (BuAer Instruction NavAer 00.55).

Incorporation of Aircraft Service Changes by operating activities will be as follows:

1. IMMEDIATE ACTION changes are to be incorporated prior to further flight.
2. URGENT ACTION changes which have an effective date are to be incorporated not later than the next intermediate inspection, or the next major inspection, with the aid of higher class maintenance activities if necessary.
3. ROUTINE ACTION changes which have an effective date are to be incorporated at the next overhaul or special modification period. When considered nec-

essary, Major Operating Commands may designate specific changes to be incorporated by activities under their control.

Changes that have been issued under the previous classifications of VICTOR, WILLIAM, XRAY, YOKE and ZEBRA and BuAer Aircraft Bulletins, will be governed by the provisions of Aviation Circular Letters 52–47 and 14–51.

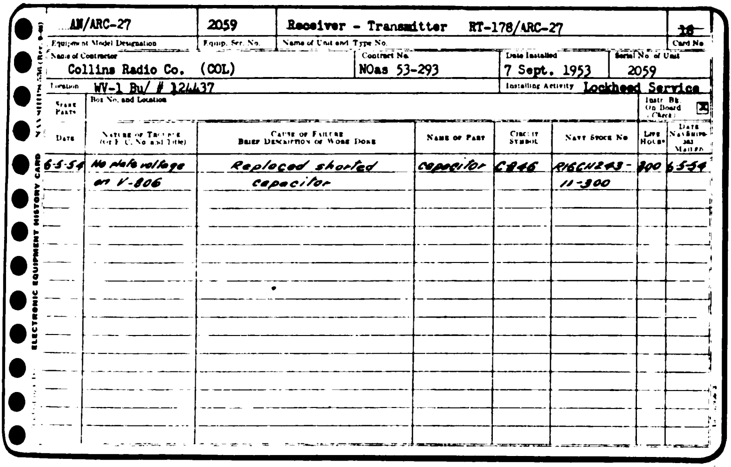
AIRCRAFT BULLETINS.—The content of these bulletins is similar to aircraft service changes. They contain only information or instructions considered essential for the satisfactory operation of specified aircraft. A bulletin should not contain instructions that involve a service change except when only a few units are involved or the change can easily be made.

ELIMINATION OF RADIO NOISE IN AIRCRAFT (AFTO 16–1–45).—The purpose of this handbook is to acquaint maintenance, communications, and engineering personnel with the types of radio noise which cause interference, and to give them a working knowledge of the techniques of noise suppression. It may be used as a guide to enable you to determine the type of interference, to locate its source, and to provide a means for its elimination or suppression.

HANDBOOK, AIRCRAFT INSPECTION REQUIREMENTS.—These handbooks contain complete requirements for periodic maintenance inspections and periodic replacement of accessories and components applicable to the aircraft to which the handbook pertains. They will prove helpful in the construction of check sheets.

REPORTS AND RECORDS

DEFECTIVE MATERIAL REPORTS.—The method of reporting defective material is determined by current BuAer Instructions. Since these change from time to time, the current method will not be presented here. Information concerning these reports (Electronics Failure Report) may be obtained from BuAer Instruction NavAer 00.58A.



ELECTRONIC EQUIPMENT HISTORY CARD

AN/ARC-27	2059	Receiver - Transmitter RT-178/ARC-27	18
Equipment Model Designation	Equip. Ser. No.	Name of Unit and Type No.	Card No.

Name of Contractor: Collins Radio Co. (COL)
Contract No.: NOas 53-293
Date Installed: 7 Sept. 1953
Serial No. of Unit: 2059
Location: WV-1 Bu/ # 124437
Installing Activity: Lockheed Service
Spare Parts — Box No. and Location:
Instr. Bk. On Board (Check) ☒

Date	Nature of Trouble (or F. C. No. and Title)	Cause of Failure Brief Description of Work Done	Name of Part	Circuit Symbol	Navy Stock No.	Live Hours	Date NavShips 383 Mailed
6-5-54	No plate voltage	Replaced shorted	capacitor	C846	R16CN243-	300	6-5-54
	on V-806	capacitor			11-300		

Figure 1–2.—History card, electronic equipment.

HISTORY CARD RECORDS.—Records are important. The more complete record you keep on your equipment the better off you will be. Different systems of recordkeeping are used by different shops; but the important thing is to be certain that complete up-to-date records are kept.

A form that may be used and is readily available is NavShips 536. This is the basic maintenance history card for electronic equipment. It provides for recording failures and other pertinent information on electronic equipments. A separate card is filled in for each equipment or major component of larger equipments. The card is so designed that when it is properly filled in necessary information is readily available for partial completion of the Electronics Failure Report. The heading of the history card should be typed, but entries on the body may be either typed or written in ink or indelible pencil. The history card is illustrated in figure 1–2.

PERFORMANCE RECORDS OF ELECTRONIC MATERIAL CHANGES.—The leading petty officers of the electronics division will be responsible for making changes to electronic equipments. The changes will be in accordance with BuAer directives. The maintenance department of your activity will receive these changes and pass them on to you in the form of work orders. As changes are made, your maintenance problems will multiply unless you effect a system of records which will allow a rapid and fully understandable picture of your equipments. A Field Change Card file system is suggested as a desirable means of keeping this information readily available. A workable form for this purpose is NavShips 537 (Record of Field Changes).

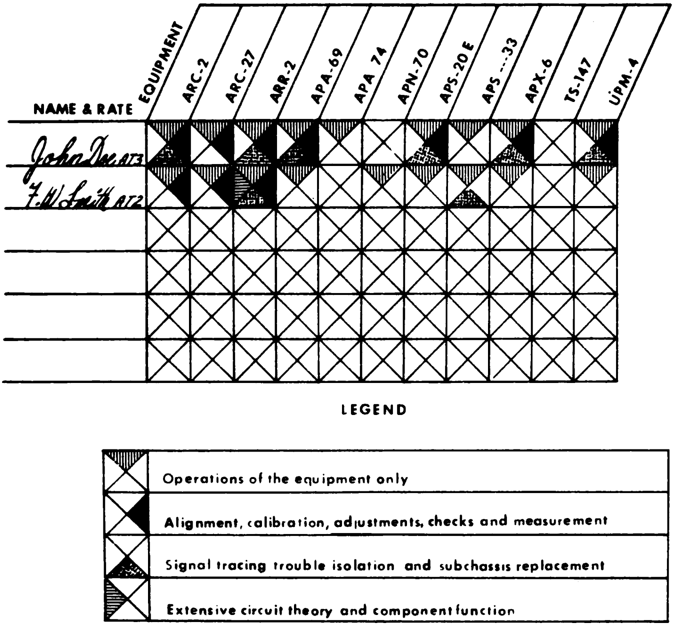
The maintenance department of your activity must be kept posted on changes that are made. This is usually accomplished by completing the work order that is issued by maintenance and these work orders are then returned to maintenance for recording.

The change or changes that are made should be noted on the equipment history card for the particular piece of gear.

WORK SHEETS AND RECORDS OF REPAIR TO AIRCRAFT.—The types of work that your division will perform will vary from jobs that require a very few minutes to those that require hours and even days. The manner in which jobs are announced will be in connection with some sort of worksheet. The sheets vary in their construction but are all essentially the same for they give necessary information relating to the work to be done.

A typical worksheet will contain the following information: authority for doing the work, work to be done, who is responsible for seeing that the work is done, when work is to be done, material to be used, and logbook entry to be made. A word of caution is in order. This type work is time consuming; it usually takes longer than you think. Form the habit of completing the work as soon as possible, otherwise you will be saying to yourself "My division is understaffed." If possible, do not allow a backlog of worksheets to accumulate.

TRAINING CHARTS AND RECORDS.—Every petty officer is charged with the responsibility of training others. The AT1 and ATC are expected to perform this phase of their work in an outstanding manner because of the nature of


EQUIPMENT
NAME & RATE
ARC-2
ARC-27
ARR-2
APA-69
APA 74
APN-70
APS-20E
APS-33
APX-6
TS-147
UPM-4
John Doe AT3
F.W. Smith AT2
LEGEND
Operations of the equipment only
Alignment, calibration, adjustments, checks and measurement
Signal tracing trouble isolation and subchassis replacement
Extensive circuit theory and component function

Note: By placing this chart behind a sheet of plexiglass and using wax pencil it is possible to replace names and equipment.

Figure 1–3.—Training chart.

the field of electronics. The work expected from those in this field is far from easy. It is a difficult area because of a number of peculiarities. Some of these are: the field is broad in its coverage; it is highly scientific; it is constantly in a state of change—new developments, new demands, new discoveries; maintenance of electronic equipment is difficult and time consuming; there is a constant change of trained personnel. These and others are reasons for you to take the job of training seriously.

Prepare for such duties by consulting with you division and education officers; they can be of much help. Study books dealing with training, organize a training program for your division, and see that the schedule is carried out.

Last, but by no means least, organize a system of recordkeeping to substantiate the training that has been given. These records are helpful in many ways. Two of the more important are: First, it lets you know what has been taught; and second, it lets the personnel know that training is important and that they are being checked as to accomplishments.

Probably the most effective and efficient means of maintaining training status is that of the training chart. These vary in construction according to needs that are served and must be made to conform to the specifications of the job to be done. The chart shown in figure 1–3 is given as an example as to the use of a training chart.

In a large organization there will be large quantities of different equipments and also many men to work with these equipments. A desirable training chart will aid the supervisory personnel in knowing which men are trained in the operation and maintenance of each piece of equipment. Figure 1–3 shows a training chart that has been found to work satisfactorily in some organizations.

TEST EQUIPMENT CALIBRATION AND REPAIR.—Test equipment must be kept in proper working order for the accuracy of other electronic gear depends on it. Inspection teams frequently ask for substantiating records that show when gear was last calibrated.

Records that are kept may be in the form of a logbook or the data may be placed on the equipment history card.

QUIZ

1. In order to meet the requirements for advancement to first class or chief, a man must have a wide background of practical experience. He should also be familiar with the requirements for advancement as listed in
   a. *Manual of Qualifications for Advancement in Rating,* NavPers 18068 (Revised)
   b. *Basic Electronics,* NavPers 10087
   c. *Aviation Electronics Technician 3 and 2,* NavPers 10317
   d. all of the above

2. One of the most important responsibilities of a first class or chief is that of
   a. supervising cleanup detail
   b. training men in lower rates
   c. regulating the workload
   d. disapproving of special liberty requests

3. To qualify for a rate you must satisfy the requirements as set forth in the
   a. *Manual of Qualifications for Advancement in Rating,* NavPers 18068 (Revised)
   b. *Bluejackets' Manual*
   c. BuAer publications
   d. *Naval Aeronautic Publications Index,* NavAer 00–500

4. Besides the professional background that is necessary for advancement to first class or chief, it is necessary to meet the military requirements that are responsibilities of increasing magnitude. The best source of information regarding military responsibilities is the publication
   a. *Basic Electronics,* NavPers 10087
   b. *Aviation Electronics Technician 3 and 2,* NavPers 10317
   c. *General Training Course for Petty Officers,* NavPers 10055
   d. *Bluejackets' Manual*

5. As a leader of men, you should not use work habits that are not in agreement with
   a. *U.S. Navy Safety Precautions,* OpNav 34P1
   b. *Naval Aeronautic Publications Index,* NavAer 00–500
   c. *Basic Electronics,* NavPers 10087
   d. *General Training Course for Petty Officers,* NavPers 10055

6. As leading petty officer, your training program should be
   a. rigid and unrelenting
   b. flexible without destroying its usefulness
   c. randomly employed
   d. all classroom work

7. An index of the efficiency of your division is
   a. the manner in which paperwork is handled
   b. the morale of the men under you
   c. neglect of chain of command procedures
   d. disregard of human relationships

8. Land-based maintenance performed by AT's is usually under which of the following type classifications?
   a. A–B
   b. B–C
   c. C–D
   d. E–F

9. Information that prescribes the replacement of electronic component parts, for example, replacing a 300-volt capacitor with a 500-volt capacitor, may be found in which of the following publications?
   a. Aircraft Bulletin
   b. EMB
   c. EMC
   d. Aircraft Service Change

10. Which of the following publications would most likely contain information pertaining to hazardous conditions in connection with testing a particular electronic equipment?
    a. Aircraft Bulletin
    b. EMB
    c. EMC
    d. Aircraft Service Change

11. A chart listing the quantities of each type of tube that each equipment uses may be found in
    a. NavAer 00–35QR–7
    b. NavAer 00–35QR–4
    c. NavAer 00–35QR–6
    d. NavAer 00–35QR–5

12. Which of the following publications would be the most desirable in helping to identify uncommon replacement parts of electronic equipment?
    a. *Handbook of Service Instructions*
    b. *Illustrated Parts Breakdowns*
    c. *ASO Catalog*
    d. Section R Allowance List

13. Complete ordering information for all publications is included in
    a. CO–NavAer 16–45–511
    b. OpNav 34P1
    c. NavAer 00–500
    d. *BuAer Manual*

14. Which of the following information on a particular tube is not given in the *Cross Index of Electron Tubes?*
    a. Stock number
    b. Types of tubes that are interchangeable with it
    c. Equipments that use the tube
    d. Types of tubes it replaces

15. The procedure to follow in making a report on defective material is prescribed in instructions from
    a. BuAer
    b. BuPers
    c. BuSandA
    d. CNO

16. In connection with test equipment, inspecting teams are frequently going to ask for substantiating records as to
    a. number of hours used each quarter
    b. types of maintenance performed
    c. number of men qualified as operators
    d. calibration

CHAPTER 2

R–C AND R–L NETWORKS

Electronic equipment that is used in present day naval aircraft is complex and everchanging. Aviation electronic equipment is used for communications, navigational aid, search, fire control, aircraft identification, and numerous other functions. The information presented in this and the following chapters is for the purpose of aiding you in gaining an understanding of these various equipments.

For the most part, the circuits discussed are common to electronic equipments that the technician will be required frequently to maintain and operate. Only in a few cases will circuits peculiar to a particular piece of equipment be used.

Many of the circuits in electronic equipment are concerned with timing and control. Circuits accomplish this function by producing a variety of voltage waveforms, such as square waves, sawtooth waves, trapezoidal waves, rectangular waves, brief rectangular pulses, and sharp peaks. In sound and in radio, tubes operate within the limits for which they are designed. In timing circuits, they often are violently overdriven, frequently operating at points which range from well in the grid current region to far beyond cutoff. Although all these circuits are broadly classified as timing circuits, the specific function of any individual circuit might be timing, waveshaping, or wave generating. This chapter discusses the principles of a variety of timing circuits, and provides a foundation for understanding the operation of specific circuits.

TRANSIENT AND NONSINUSOIDAL VOLTAGES

Most of the waveshapes in timing circuits are produced by combinations of vacuum tubes, resistors, capacitors, and inductors. For example, the charging and discharging times of capacitors and inductors make these elements

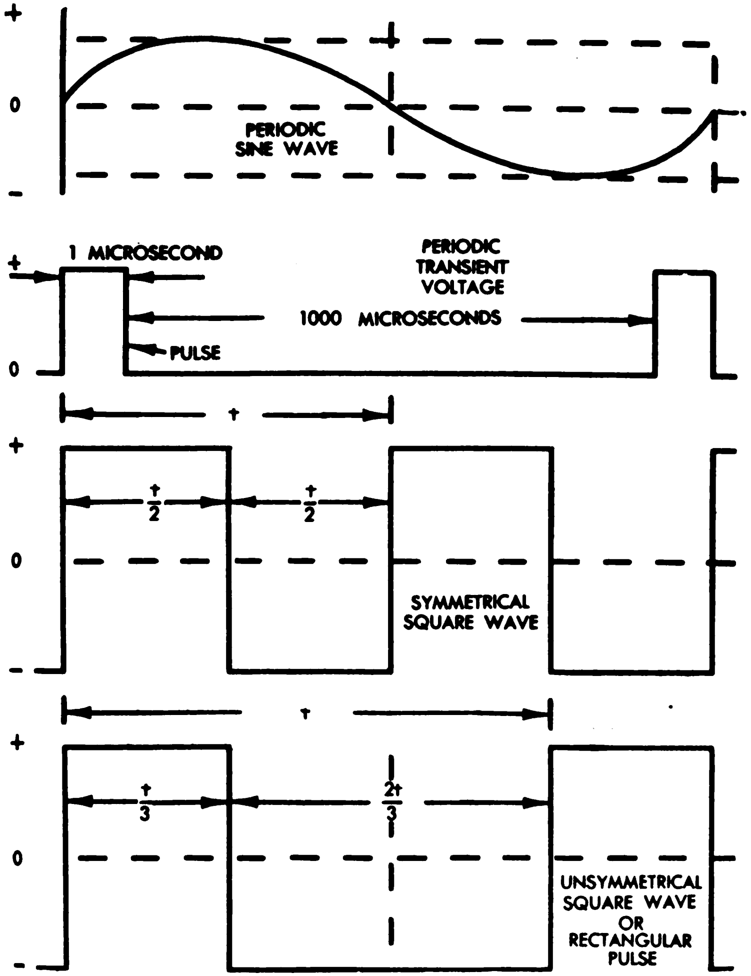


Figure 2–1.—Typical transient voltages.

extremely useful in timing circuits because of their effect on transient and nonsinusoidal voltages.

A transient nonsinusoidal voltage is a brief voltage change called a pulse which occurs once and does not recur for a comparatively long time. Such a voltage change is never sinusoidal but is usually periodic; that is, the change repeats itself at regular intervals. Examples of typical transient voltages are compared to a periodic sine wave in figure 2–1.

Analysis

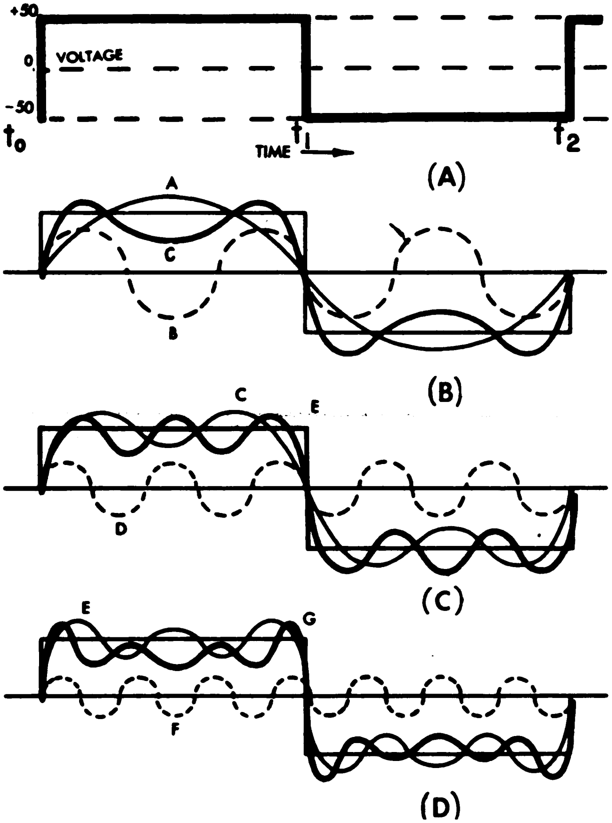
For the purpose of circuit design, there are two ways of analyzing transient nonsinusoidal voltages. One is to consider that the transient waveshape is merely a rapid change of voltage, which may be followed after a certain interval by another and similar change. The other is to assume that the waveshape is the result of the addition of a large number of sine waves having different frequencies and amplitudes. The second method of analysis is more useful in most cases. In the design of an amplifier, for example, this type of analysis is necessary because the amplifier cannot handle a transient without distortion unless it is capable of handling all the sine-wave frequencies making up the transient.

Transient voltages have many different and complex shapes. The more commonly used transient waves are the square wave, the sawtooth or triangular wave, and the peaked wave.

The Square Wave

Figure 2–2 (A) shows a square wave in which voltage is plotted against time. This wave is called SYMMETRICAL because the alternations are identical.

In terms of the first method of analysis stated in the preceding paragraph, you can consider the square wave in the illustration as a voltage which remains unchanged at plus 50 volts until time t_1, when it suddenly changes to minus 50 volts; it remains at this value until time t_2, then suddenly rises to plus 50 volts and remains at this value until time t_3, and so on.



A FUNDAMENTAL
B 3RD HARMONIC
C FUNDAMENTAL PLUS 3RD HARMONIC
D 5TH HARMONIC
E FUNDAMENTAL PLUS 3RD AND 5TH HARMONIC
F 7TH HARMONIC
G FUNDAMENTAL PLUS 3RD, 5TH AND 7TH HARMONIC

Figure 2–2.—Composition of symmetrical square wave.

In terms of the sine-wave method, you can analyze the square wave by determining what sine waves are required to reproduce it. To reproduce a symmetrical square wave, you start with a sine wave, which is equal to the square wave repetition frequency, and add to it the odd harmonics of this frequency as shown graphically in (B), (C), and (D) of figure 2–2. The waveshape *C* is formed by adding the fundamental frequency *A* and its third harmonic *B* which has an amplitude one-third that of the fundamental. (See fig. 2–2 (B).) Notice that the resultant wave already slightly resembles a square wave as you can see by the square wave superimposed on the diagram. The result of adding the 5th harmonic of the fundamental frequency to the resultant wave *C* is shown in figure 2–2 (C). Notice in the resultant wave *E* that the corners are much sharper and the top somewhat flatter.

The 7th harmonic at 1/7 amplitude is added to form the resultant curve *G* (fig. 2–2 (D)). This curve is fairly smooth across the top and fairly sharp at the corners. Adding the 9th harmonic at 1/9 amplitude, the 11th harmonic at 1/11 amplitude, etc., would further sharpen the corners and flatten the top of the wave.

The assumption that a square wave contains a number of frequencies, although it is actually just a change of voltage, is useful in studying the square wave in circuits. For example, according to the frequency method of analysis the higher harmonics are responsible for the sharpness in the corners of the square wave. Therefore, rounding off the corners of the square wave can be explained as discriminating against high frequencies by the circuit. Furthermore, since this method of analysis reduces a square wave to terms of sine waves of different frequencies, it is possible for engineers to design circuits by using conventional sine-wave engineering practices.

Most square waves employed in electronics consist of short pulses separated by long time intervals. (See fig. 2–3 (A).) Such pulses are constructed in the same way as square waves, but in addition to the sine waves of the fundamental pulse frequency and many harmonic frequencies of fractional amplitudes, a small d-c voltage is required. The duration of the pulse is one-third of the repetition time which, in this case, is 1,000 times per second (fig. 2–3 (A).) To reconstruct this pulse, add

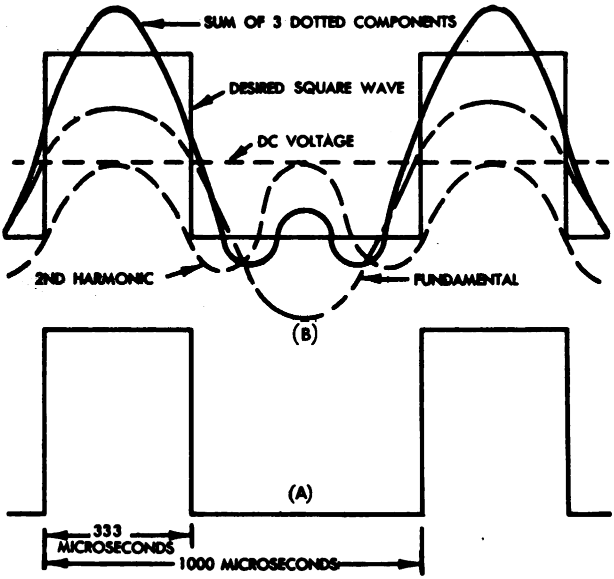


Figure 2–3.—Composition of square wave with short duration pulses.

a d-c voltage equal to one-third the desired pulse amplitude to a 1,000 c.p.s. sine wave having an amplitude two-thirds of the pulse amplitude, a 2,000 c.p.s. sine wave having an amplitude one-third of the pulse amplitude, and so forth. The solid curve (fig. 2–3 (B)) represents the sum of all these voltages. Notice that there is a rough resemblance between the outline of the solid curve and the desired square wave. If you add all the harmonics up to about the 15th harmonic (15,000 c.p.s.) in the proper phase and amplitude, the resultant waveshape will be quite square.

An important fact to consider is the number of harmonics to include in forming the waveshape. The rule to follow is that the maximum number varies inversely with the pulse width. This means the wider the pulse, the smaller the number of harmonics, and inversely, the narrower the pulse, the greater the number of harmonics. For example, in figure 2–3 the width of the pulse is such that the first 15 harmonics can be included.

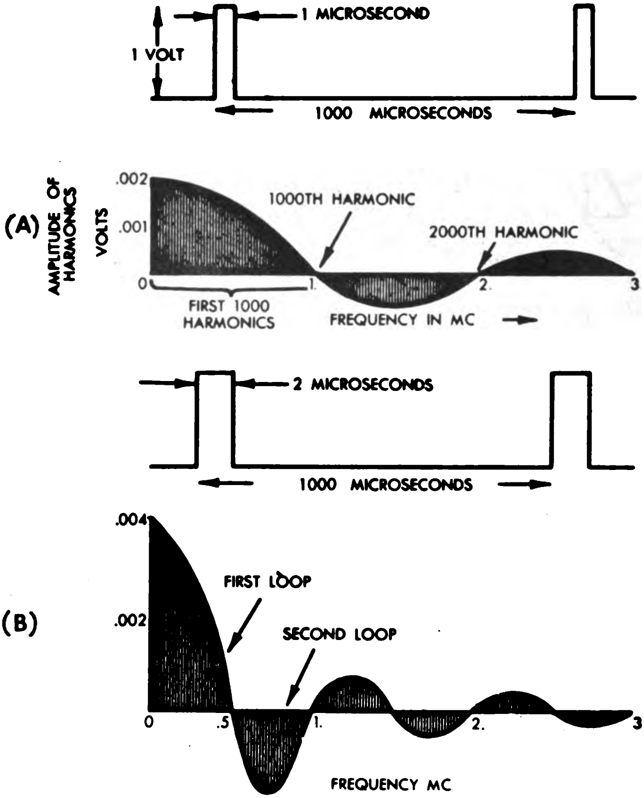


Figure 2–4.—Harmonic content of short square pulses.

In figure 2–4 a much narrower pulse is shown, a one-microsecond square wave having a repeating frequency of 1,000 c.p.s. This pulse requires a great many more harmonics, as you can see in (A) of figure 2–4. Here, the amplitude of each harmonic is plotted on the *Y* axis, and its frequency is plotted on the *X* axis. The tops of all the vertical lines are connected by a curve which forms a sine wave. It is impossible to show all the frequencies, since 1,000 harmonics actually occur between zero and 1 megacycle.

When the pulse amplitude, as shown in figure 2–4, is one volt, the damped sine wave produced by connecting the amplitudes of the harmonics is less than 0.002 volt. The relative importance of the various frequencies depends upon the area under the sine-wave curve, varying with the size of the area. Notice that the largest area occurs between zero and 1 mc. The significance of this is that any circuit which does not discriminate against any frequency from 1,000 c.p.s. to 1 mc. will pass this pulse with little distortion.

Figure 2–4 (B) shows that doubling the pulse width to two microseconds causes the amplitude of the harmonic frequencies to drop to zero at a half megacycle. Thus, any circuit designed for operating with a two-microsecond width will not discriminate against any frequency from 1,000 c.p.s. to 500,000 c.p.s.

All this leads to the conclusion that there is a certain relationship between the highest frequency that a circuit can handle and the width of the pulse. For example, in the case of the one-microsecond pulse, the highest frequency was 1 mc., and in the case of the two-microsecond pulse, 1/2 mc. The rule is that frequency varies inversely with pulse width. Mathematically this is expressed as,

$$F_h = \frac{1}{PW}$$

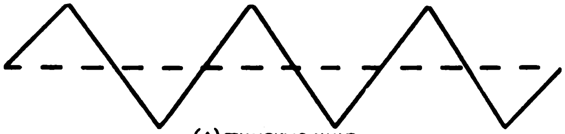
where F_h is the highest frequency for the circuit, and PW the pulse width in microseconds when the pulse width is much smaller than the pulse repetition period.

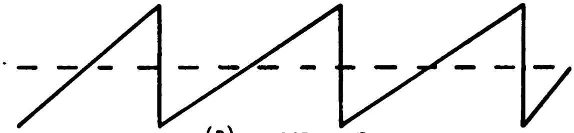
Note that the pulse repetition frequency is not considered in the equation, since it is quite evident also by transient analysis that the pulse recurrence frequency has nothing to do with the harmonic frequencies in the pulse. For example, if the repetition frequency changed to 500 c.p.s. (fig. 2–4 (B)), there would be twice as many harmonic frequencies in the first loop, but the loop would still pass through zero at 0.5 mc.

The lowest frequency that this circuit must be able to handle is the fundamental or the pulse repetition frequency. However, several of the low-frequency harmonics can be removed from the curves without noticeably reducing the areas. (See fig. 2–4 (A) and (B).) Thus, a circuit may be designed for a one-microsecond pulse and handle a frequency band as narrow as 10 kc. to 1 mc.

Triangular and Sawtooth Waves

Triangular and sawtooth waves, like square waves, can be constructed with a series of sine waves. Circuits designed to handle the short sawtooth wave shown in (C) of figure 2–5 must have about twice the bandwidth as a circuit required for a square pulse of the same dimensions.


(A) TRIANGULAR WAVE


(B) SAWTOOTH WAVE

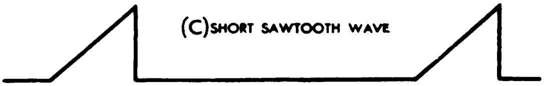

(C) SHORT SAWTOOTH WAVE

Figure 2–5.—Triangular and sawtooth waves.

COMPOSITION.—A sawtooth wave contains both odd and even harmonics. The wave shown in (D) of figure 2–6 contains all the harmonics up to the 7th. Notice how closely this wave resembles the sawtooth wave superimposed on it.

From the viewpoint of a transient voltage, you can consider the sawtooth voltage wave a slow linear voltage change from an initial value to a certain higher value, followed immediately by a sudden drop to the initial value. The wave may repeat immediately or after an interval of time has elapsed.

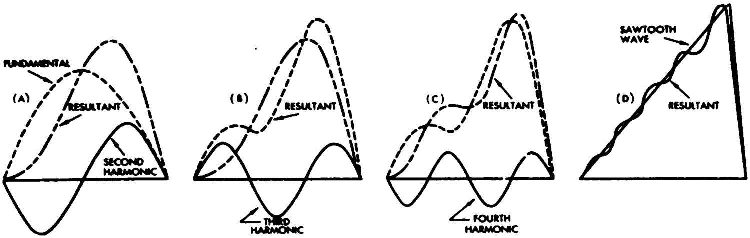


Figure 2–6.—Composition of sawtooth wave.

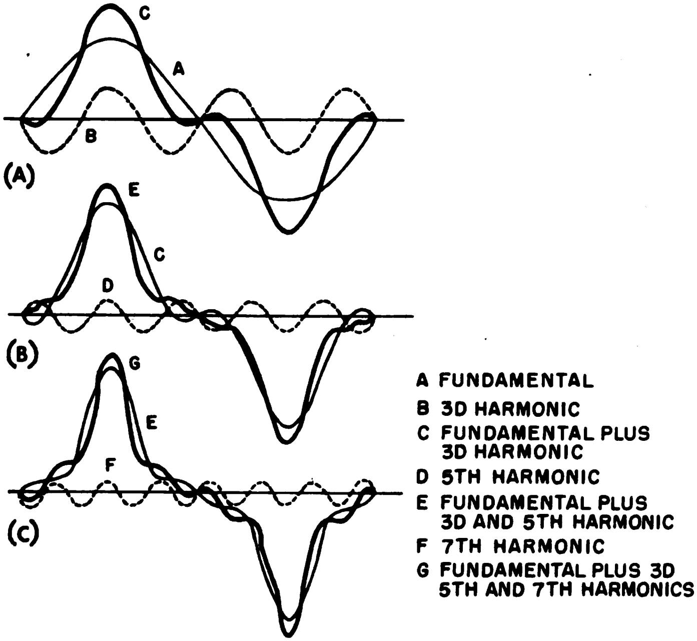

C
A
B
(A)
E
C
D
(B)
G
E
F
(C)
A FUNDAMENTAL
B 3D HARMONIC
C FUNDAMENTAL PLUS 3D HARMONIC
D 5TH HARMONIC
E FUNDAMENTAL PLUS 3D AND 5TH HARMONIC
F 7TH HARMONIC
G FUNDAMENTAL PLUS 3D 5TH AND 7TH HARMONICS

Figure 2–7.—Composition of peaked wave.

Peaked Waves

A peaked wave is another wave which can be constructed with a number of sine waves. Figure 2–7 shows how the addition of successive odd harmonics to a fundamental sine wave produces a peaked wave. First, the 3rd harmonic, a wave having a lower amplitude, is added to the fundamental. Next, the 5th harmonic is added to wave *C*. This produces wave *E* in which the peak is further heightened and the sides further sloped. This process, if carried further, will produce a wave with higher and higher peaks and steeper and steeper sides.

The composition of a peaked wave differs from that of a square wave in that there is a different phase relationship between the harmonics. In the square wave composition all the odd harmonics cross the zero line in phase with the fundamental. In the peaked wave composition the 3rd, 7th, 11th, and so forth, harmonics cross the zero line 180 degrees out of phase with the fundamental while the 5th, 9th, 13th, and so forth, cross the zero line in phase with the fundamental.

TRANSIENT VOLTAGES IN R–C CIRCUITS

R–C Time Constant

The characteristics of the waveforms required by special circuits, such as slope, duration, or repetition frequency, are determined largely by controlling the variation of a voltage with respect to time. Since the charging or discharging of a capacitor through a resistor requires time, resistance-capacitance circuits are commonly used for this purpose.

To learn how a resistor-capacitor circuit affects a waveshape applied to it, it is well to investigate first the voltage and current changes which occur when a d-c voltage is applied to the circuit. Most R–C circuits and the voltages applied to them are reducible to a simple equivalent circuit containing a battery, capacitor, and resistor.

Figure 2–8 (A) shows such a simple series circuit. According to capacitor theory, the capacitor C will offer zero resistance to the direct-current flow in the circuit at the instant the switch is closed. Later the capacitor C will offer infinite resistance, since it charges to a voltage equal and opposite to the battery voltage E_{bb} after a period of time. When the capacitor is charged, the battery voltage is cancelled completely, leaving no voltage to cause a current to flow through the resistor.

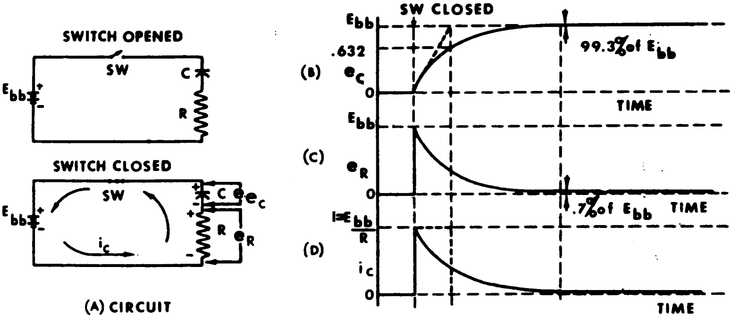


Figure 2–8.—Current and voltage during capacitor charge.

However, before the current does reach zero, there is a lapse of time. This lapse provides the time element required in the formation of many different waveshapes.

To understand how a capacitor affects the time delay in a circuit, first consider what happens in the capacitor itself, and how it is affected by the rest of the circuit. Assume that before you close switch *Sw*, the capacitor is completely discharged—that is, there is an equal number of electrons on each of its plates. When you close the switch, the battery voltage E_{bb} immediately forces electrons into the capacitor plate connected to the negative terminal of the battery, placing a negative charge on it. This creates a stress on the dielectric which, in turn, forces electrons away from the other plate and makes its charge positive.

It might seem that if it were not for the limiting effect of the series resistor, the number of electrons that would accumulate on the capacitor plates would be infinitely high because, upon the closing of the switch, the current in the circuit would immediately jump from zero to a value determined only by the resistor. (That is, $i=\frac{E_{bb}}{R}$ since the uncharged capacitor offers no opposition to current flow.)

The fact is, however, that the capacitor itself offers opposition to the current. Just as soon as a few electrons accumulate on the capacitor plate, there is a like number forced away from the other plate. The net result is a small voltage difference across the capacitor, the plate with more electrons being the negative point of voltage, and the plate with less electrons being the positive point of voltage. The polarity of the circuit then becomes as shown in (A) of figure 2–8. Notice here that the capacitor is charged to a voltage opposite in polarity to the battery voltage. Thus, in a short time after you close the switch, the movement of electrons—that is, current flow—does not remain at its initial value, but drops to a lower value because the voltage across the capacitor subtracts from the charging voltage. Since the voltage applied to the capacitor is thus effectively reduced, it charges at a lower rate.

Now consider the action resulting during the further charge at a slow rate up to the point of so-called complete charge. (Theoretically the capacitor never completely charges to 100 percent of the applied voltage.) When the switch was first closed (fig. 2–8 (A)), the capacitor charged rapidly to 63.2 percent of the applied voltage. At this point the current flow in the circuit was 36.8 percent of its initial value. With the continuous increase in the voltage difference across the capacitor and its subtraction from the voltage applied to it, the charge rate tapers off, but the charge itself rises exponentially to the point where the capacitor is considered charged.

Notice the voltage and current curves shown in (B), (C), and (D) of figure 2–8. Had the capacitor charged at its original rate, its charge curve would be the dotted line in (B). The slope of this capacitor charge curve is expressed as $\frac{E_{bb}}{RC}$.

The term RC in this ratio is called the time constant (*t.c.*) of an R–C circuit. It is the time required to charge a capacitor to 63.2 percent of its final voltage. To find the time constant in seconds, multiply R, the resistance in ohms, by C, the capacitance in farads. When using microfarads for C, the time constant will be in microseconds.

The capacitor voltage curve in (B) of figure 2–8 and the instantaneous current curve in (D) are exponential curves. An exponential curve is the graph of an equation in which the independent variable appears as an exponent.

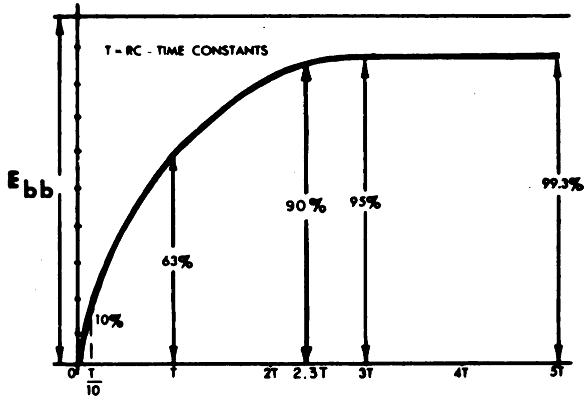
The instantaneous resistor voltage curve in (C) of figure 2–8 can be found from the instantaneous current values by using the Ohm's law formula $E=IR$. Since R is constant, E, the voltage across the resistor, is directly proportional to the current through it. Therefore, its curve has the same exponential slope as the instantaneous current curve.

Values of Capacitor and Resistor Voltages

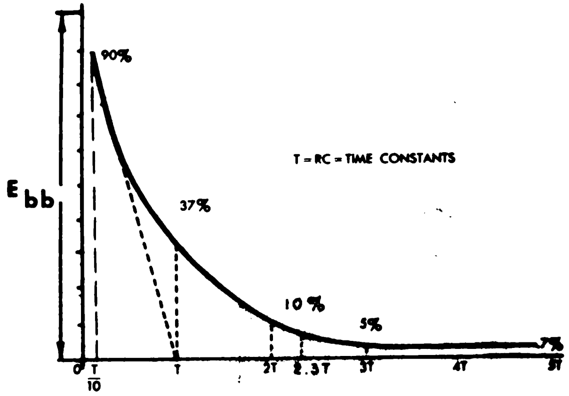
There are three methods of determining the instantaneous values of voltage across capacitors and resistors in R–C circuits. The first is a rough estimate from the exponential curve with the values at four main points known. The second is a close approximation from a very accurate exponential curve on a universal time constant chart. The third involves solving the equation for the curve with the quantities for the R–C problem inserted in the equation.

ROUGH APPROXIMATION CURVES.—The voltage or current at any instant during the charging time of a capacitor can be approximated from the two curves shown in figure 2–9. These curves employ only rough values, but it is well that you memorize them, since they occur frequently. They are also useful for approximating current at any instant during the charge of a capacitor. To find current, substitute the total current flow in the circuit for E_{bb} on the curves.

UNIVERSAL TIME CONSTANT CURVES.—A universal time constant chart (fig. 2–10) provides a much more precise method for determining the voltage and the current at any instant during the charge of a capacitor than the



(A) CURVE FOR CAPACITOR VOLTAGE DURING CHARGE



(B) CURVE FOR RESISTOR VOLTAGE DURING CHARGE

Figure 2–9.—Exponential curves for rough approximations.

method just described. The chart shows a number of time constants plotted against the fraction of maximum current or voltage.

In learning to use this chart study the following examples:

Problem 1.—In the circuit in figure 2–11, assume that C equals 0.001 μf. and that R equals 100K. Find the capacitor voltage 200 microseconds after the switch is closed.

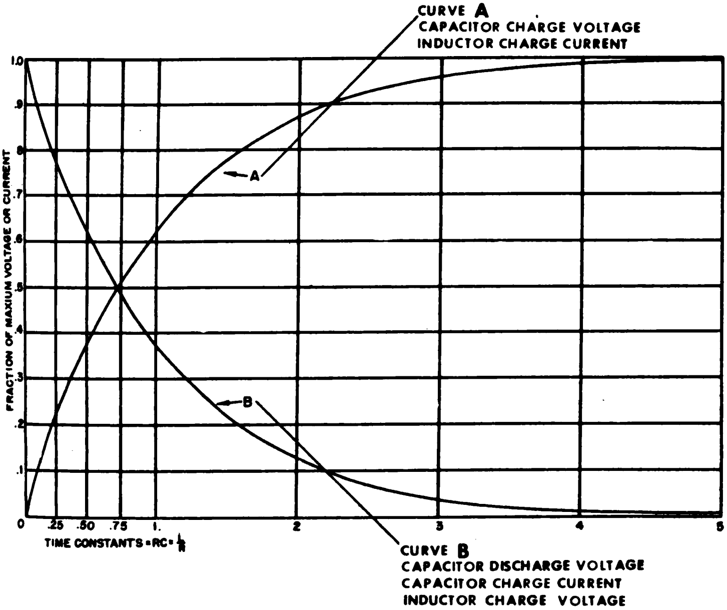


Figure 2–10.—Universal time constant chart.

Solution:

a. Find the time constant.

$$t.c. = RC$$
$$t.c. = 100{,}000 \times 0.001$$
$$t.c. = 100 \text{ microseconds}$$

b. Since 200 microseconds is equivalent to 2 time constants, enter the universal time constant chart at 2 along the bottom and proceed up to curve *A*. (This curve represents the capacitor charge voltage.) At curve *A*, proceed horizontally to the vertical scale and read 0.86. This represents the fraction of maximum voltage.

c. Thus, the capacitor voltage is about 86 percent of maximum. With 200 volts as the applied voltage, the capacitor voltage equals

$$200 \times 0.86 \text{ or } 172 \text{ volts.}$$

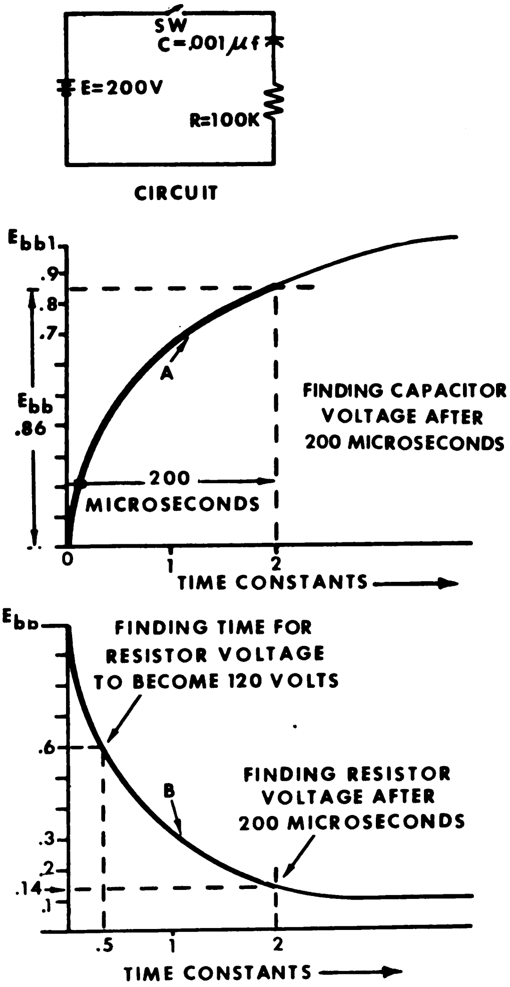


Figure 2–11.—Using universal time constant chart.

Problem 2.—In the same circuit find the resistor voltage 200 microseconds after the switch is closed.

Solution, using Kirchhoff's law:

From the previous example, the capacitor voltage equals 172 volts.

Since the circuit is a series circuit, the resistor voltage is the difference between the applied voltage and the capacitor voltage. Thus,

$$200-172=28 \text{ volts.}$$

Solution, using the universal time chart:
Knowing that the resistor voltage starts high and then drops off, you use curve *B*. Since there are two time constants involved, go up from 2 to curve *B* and then across the vertical scale. At the intersection you find 0.14. Since the applied voltage is 200 volts, the resistor voltage must be

$$200\times 0.14 \text{ or } 28 \text{ volts.}$$

Solution:
Using universal chart; find current.
By Ohm's law the maximum current is

$$\frac{200}{100K} \text{ or } 2 \text{ ma.}$$

At two time constants, the current is 0.14 of maximum, according to curve *B* and the scale at the left. Therefore,

$$0.14\times 2 \text{ ma.}=0.28 \text{ ma.}$$

By Ohm's law

$$E=IR$$

The instantaneous voltage is

$$0.28 \text{ ma.}\times 100K=28 \text{ volts.}$$

Problem 3.—In the same circuit find how long it takes for the resistor voltage to drop to 120 volts.

Solution:

$$120 \text{ volts}=\frac{120}{200} \text{ or } 0.6 \text{ of } 200 \text{ volts.}$$

To find 0.6 on the vertical scale, go to curve *B*, thence down to 0.5 time constants. If one time constant equals 100 microseconds, the time required for the resistor voltage to drop to 120 volts must equal

$$0.5\times 100 \text{ or } 50 \text{ microseconds.}$$

NOTE: Any value of voltage and any size of resistor or capacitor can be used with the universal time constant chart. After 5.0 time constants the capacitor is considered to be fully charged, and the resistor voltage and the circuit current are considered to be zero as shown in figure 2–11.

Capacitor Discharge Through Resistance

The discharge of a capacitor through a resistor follows the same exponential curve as the charge through a resistor. (See fig. 2–12.) When the capacitor is charged to a voltage of E_c, there is an excess of electrons on one plate of the capacitor. This unbalanced condition causes a potential difference which exists as long as the switch is open (assuming the capacitor is perfect). The portions of the curves indicating an open switch condition show that the capacitor voltage is E_c, the current in the circuit is zero, and the resistor voltage is also zero.

When the switch is closed, the electrons on the negative plate immediately surge through the circuit toward the other capacitor plate, building up a current which is limited to a value determined by Ohm's law; that is, a value equal to $\frac{E_c}{R}$. This is what happens. The initial current causes a voltage drop across the resistor equal to the capacitor voltage. But as soon as current flows, the number of excess electrons causing the current flow decreases. This causes a voltage decrease which in turn decreases the current. The decrease is not linear, however. The high initial current decreases the voltage rapidly, but subsequent lower voltages cause lower currents, which remove the charge at a lower rate. The discharge becomes slower and slower until at the end of 5 time constants, the current is less than 1 percent of the initial value. The voltage causing this low current is also extremely low because the capacitor is 99.4 percent discharged. Theoretically, the time required to discharge the capacitor is infinitely long, but for practical purposes, 5

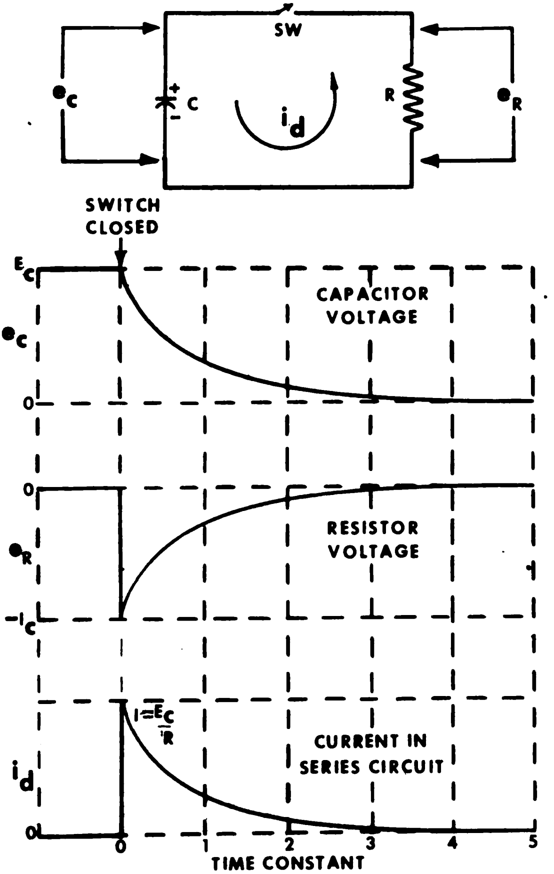


Figure 2–12.—Current and voltage during capacitor discharge.

time constants may be considered as the time required for complete discharge.

SOLVING CAPACITOR DISCHARGE PROBLEMS.—To determine instantaneous values of current and voltage in R–C circuits during capacitor discharge, use the same methods that you used previously in the charging circuit problems. One fact of significance in discharging circuits is that the voltages across the resistor and the capacitor follow the same curve inasmuch as they both start at high values and decrease exponentially to zero. Therefore, curve *B* of the universal time constant chart represents the instantaneous values of voltage across both the capacitor and the resistor during capacitor discharge. The only requirement is that e_R be assigned a polarity opposite that of e_C. Remember that the voltage across the resistor is not only equal but opposite to that across the capacitor when the capacitor is discharging. This is as it should be according to Kirchhoff's law, which states that the sum of the voltages around a series circuit is zero.

As just mentioned, you can use either method given earlier to solve for the voltage and current values; that is, the curves which roughly approximate voltages and currents or the universal curves.

Examples:

Problem 1.—Assume that in the circuit of figure 2–13 the capacitor is charged to 200 volts when the switch is closed. Find the capacitor voltage 300 microseconds after the switch is closed.

Solution:

First, find time constant by the equation

$$t.\ c.=RC.$$

Substituting the values indicated in the circuit,

$$t.\ c.=100\text{K}\times 0.001\ \mu\text{f}.$$

$$t.\ c.=10^5\times 10^{-9}=10^{-4}\text{ seconds}$$

Therefore,

$$t.\ c.=100\text{ microseconds}.$$

Since the time constant is 100 microseconds and since the problem requires 300 microseconds, the capacitor discharged for 3 time constants. Refer to the heavy lined curve shown in (B) of figure 2–9 or (B) of figure 2–13. There you see that at 3 time constants the capacitor has discharged to 5 percent of E_c or 0.05×200, or 10 volts. Thus, after 300 microseconds, the capacitor voltage is 10 volts.

Problem 2.—Find the current 200 microseconds after the switch is closed.

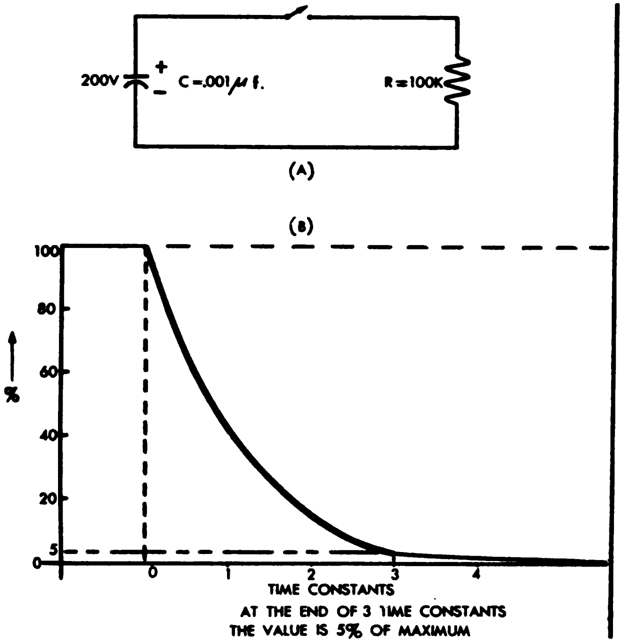


Figure 2–13.—Capacitor discharge problem.

Solution:

The current decreases from an initial value of 2 ma. (calculated by Ohm's law, $\frac{E}{R}=\frac{200\text{V}}{100\text{K}}$). The time of discharging is 200 microseconds, or 2 time constants. Referring to the universal time constant curve *B* (fig. 2–10), the value of current at 2 time constants is 14 percent of 2 ma. or 0.28 ma.; the current after 200 μsec. is 0.28 ma.

Problem 3.—Find the resistor voltage after one time constant, using time constant curve *B*.

Solution:

After 1 time constant, the voltage drops to 37 percent of its maximum value. Since the resistor voltage starts at 200 volts and drops to zero when the capacitor is completely discharged, it is 37 percent of 200 volts, or 74 volts, 1 time constant after discharge starts.

Notice here that curve *B*, which is the capacitor discharge voltage curve, also is used for voltage decrease across the resistor during discharge.

Effect of Varying *R* and C

Thus far, the R–C curves illustrated were drawn separately, but in figure 2–14 three curves are superimposed on one another for the purpose of showing what effect changing the time constant (*RC*) has on the waveshape produced by an R–C circuit. Each curve represents an R–C circuit having the same applied voltage but a different time constant. Naturally, in all R–C circuits, values reach maximum (or zero) after 5 time constants. However, since the duration of each time constant is different, the curves differ in duration. The curve labeled 20 microseconds, for example, is produced by a small R–C; that is, a small capacitor or resistor. A small capacitor has the property of quick charging, and a small resistor connected to it permits a high charging current to flow, further facilitating rapid charging. Due to the small R–C, this curve reaches its maximum limit quickly or in approximately 100 microseconds.

Larger *R–C* values produce longer time constants. A larger capacitor requires more current to charge it completely, while a larger resistor limits the current more and thereby increases the time required for charging. Thus, the curve labeled 200 microseconds has a longer time constant. Notice that 1,000 microseconds are required to fully charge this capacitor.

Notice that the e_R and i curves both follow the capacitor charge curve very closely. The first half of each of these curves represents the capacitor charge, and the last half the capacitor discharge. The difference between the two halves is due to the position of the switch.

LONG AND SHORT *RC* TIME CONSTANTS.—In contrast with the time constants shown in previous illustrations, the time constant used in figure 2–15 is called a MEDIUM

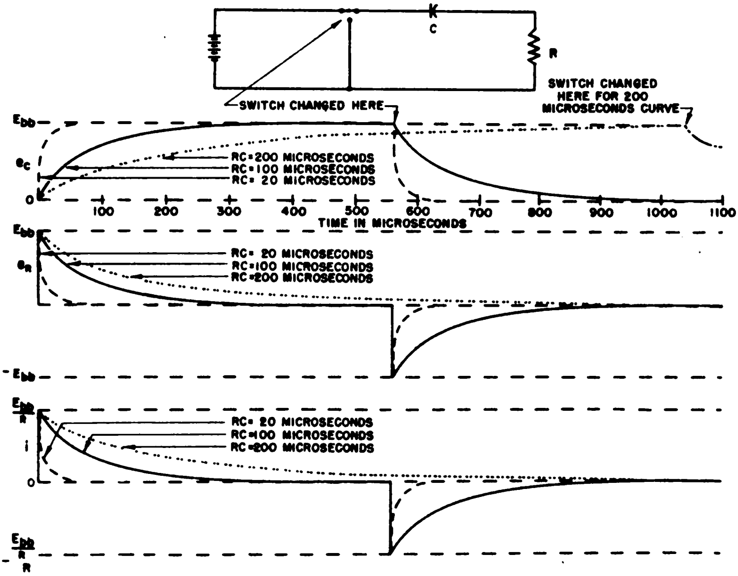


Figure 2–14.—Effect of varying *R* and *C* in R–C circuit.

TIME CONSTANT. Actually any distinction between long, short, or medium time constants is arbitrary, but in general the practice is to consider that a circuit has a long time constant when the *RC* product is equal to or greater than ten times the period of the applied voltage, a short time constant when the *RC* product is equal to or less than one-tenth the period of the applied voltage, and a medium time constant when the *RC* product lies between these two limits. However, after five *RC* seconds the charge is about 99.5 percent. For practical purposes the capacitor may be considered to be fully charged at the end of this time. When a capacitor is discharging, the discharge rate is the same as the charge rate; that is, it discharges 63.2 percent of its remaining charge in each *RC* time constant. After five *RC* seconds the capacitor can be considered to be completely discharged.

It should be remembered that when determining whether the time constant is long or short the period of the applied voltage refers to either the charge or discharge period and not to a combination of the two. For example, figure 2–4 (A) shows an unsymmetrical square wave in which the period of the applied voltage is one microsecond. In determining the action of an R–C circuit on the pulse illustrated in this figure the period would be taken as one microsecond and not 1,000 microseconds (time between pulses) in order to determine whether the *RC* is long or short.

EFFECT OF A MEDIUM TIME CONSTANT.—Since most R–C circuits in electronics are associated with high-speed timing devices, the voltages generally applied to an R–C circuit are rapidly repeating square wave voltages rather than d-c voltages. Figure 2–15 shows the capacitor and resistor voltages resulting from applying a square wave having a period of 2.5 time constants to the circuit at the top. For the purpose of analyzing the circuit, the square wave generator is redrawn into two equivalent circuits. They make it easier for you to understand the action of the generator, providing that you regard the square wave first as a constant voltage lasting for a definite time, and then as a zero voltage lasting for an equal time. Since the constant voltage equals the battery voltage during the first half of the square wave, the equivalent circuit for this time period (t_1) shows the battery voltage. During the second half of the square wave, or the zero voltage period, the capacitor discharge current flows through the generator. It is therefore not necessary to include the battery in the (time t_2) equivalent circuit. The battery, however, must be replaced by whatever resistance the discharge current encounters in the generator, which in the case being considered is zero resistance.

EFFECT OF SHORT TIME CONSTANT.—Consider now the effect of a short time constant on an R–C circuit. The square wave shown in figure 2–16 is applied to a resistance-capacitance combination with a time constant equal to 0.02 of a cycle. Thus, in 20 percent of a half cycle, five time constants will occur. If you assume as before that the square wave is a d-c voltage applied for a short time

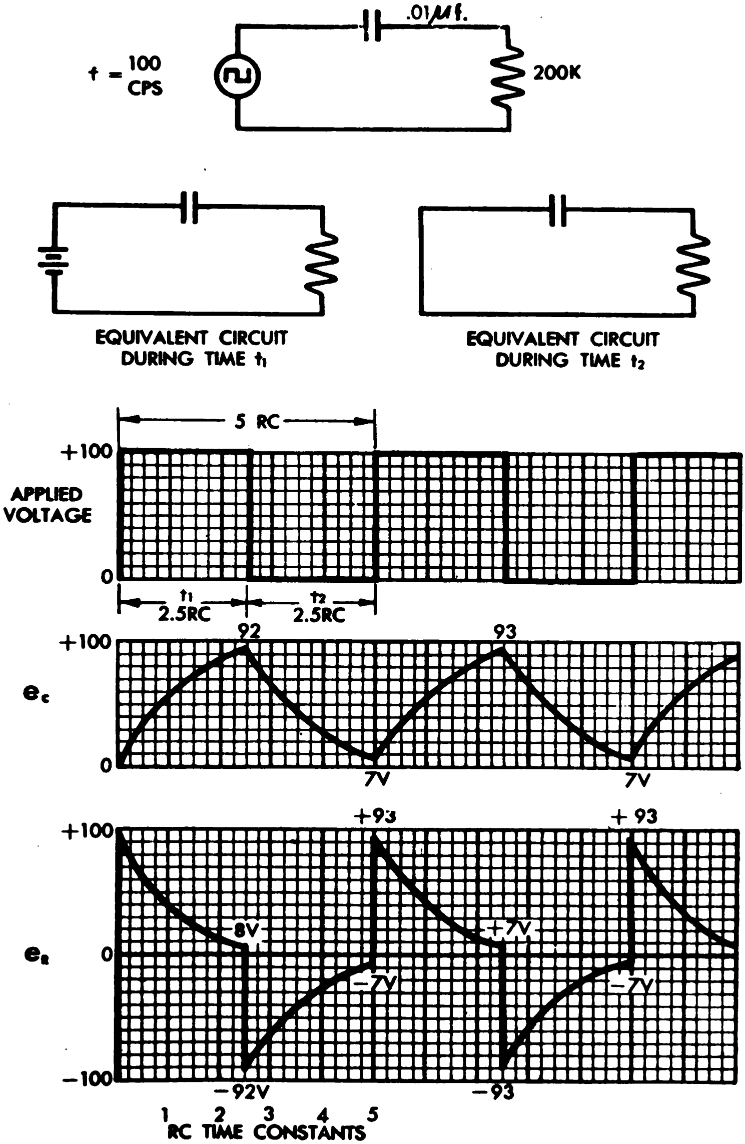


Figure 2–15.—Effect of medium time constant on square wave input.

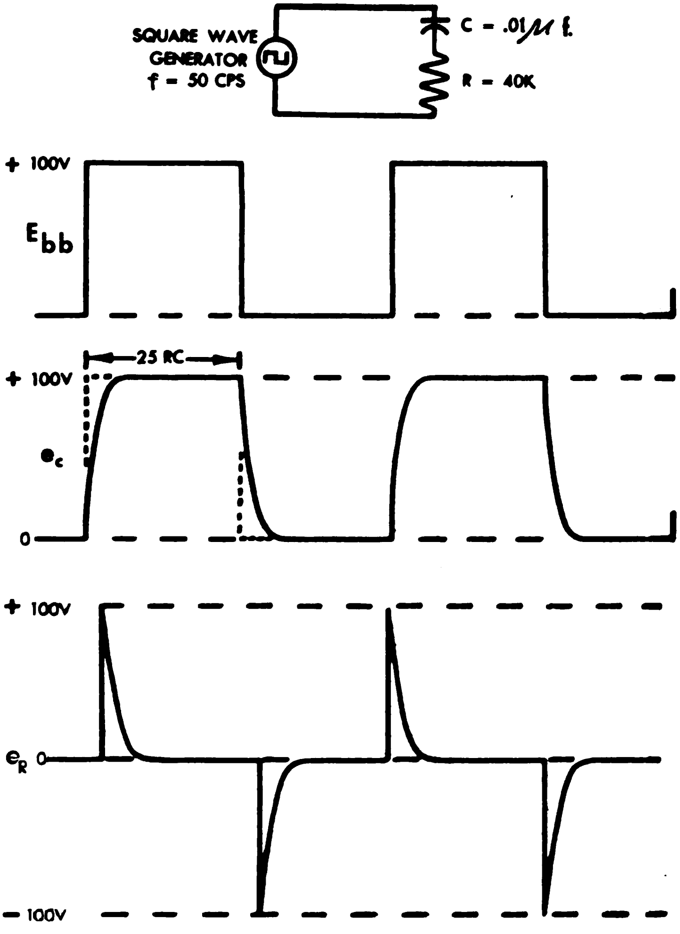


Figure 2–16.—Waveshapes for short time constant R–C circuit.

and then shorted out, you can apply the rules you used previously. These rules stated that the charging rate for a capacitor with a short time constant is very high at first and then drops slowly to zero. Similarly, its discharge rate is very high to start with and then gradually decreases to zero. The early rise of the capacitor voltage to the full applied voltage and its quick drop to zero cause the capacitor waveshape to resemble the square wave input. On the resistor the short duration of the high charging and discharging currents have a quite different effect. The rapid rise and drop of the resistor voltage cause the resistor waveshape voltage to be peaked at each square wave variation (change from maximum voltage to zero or from zero to maximum). These sharp peaks are excellent for initiating action in another circuit, as, for example, in a radar transmitter which is designed to transmit a one-microsecond pulse whenever it is triggered by a positive voltage. The sharp positive pulses shown in the illustration may be used for this purpose since they occur at precisely the changeover time in the square wave.

Referring again to figure 2–16, notice that the universal exponential curve occurs over and over in the waveshapes. You can readily determine the time and amplitude of each curve from the universal time constant curve.

EFFECT OF A LONG TIME CONSTANT.—Figure 2–17 shows the waveshapes produced by an R–C with a long time constant; the time constant is ten times as long as the period. On referring to the universal curve, you will see that the capacitor in the circuit will charge almost linearly for about 10 percent of a time constant. This causes the capacitor curve in the case illustrated to be linear during the entire half cycle. The dotted curve shows how the capacitor would have continued to charge if the applied voltage had not been removed. Note how the voltages in the circuit illustrated comply with Kirchhoff's law which states that the sum of the voltage-drops around the circuit must equal the applied voltage. During the first half cycle, the applied voltage is 100 volts. Although the resistor and capacitor voltages are constantly changing, their sum at any one instant is also 100 volts. Thus, at the end of the first half cycle, for example, the capacitor voltage is 20 volts, and the resistor voltage is 80 volts, making a total of 100 volts and fulfilling Kirchhoff's law. During the other half cycle, the applied voltage from the generator is zero. In this case the only voltage in the

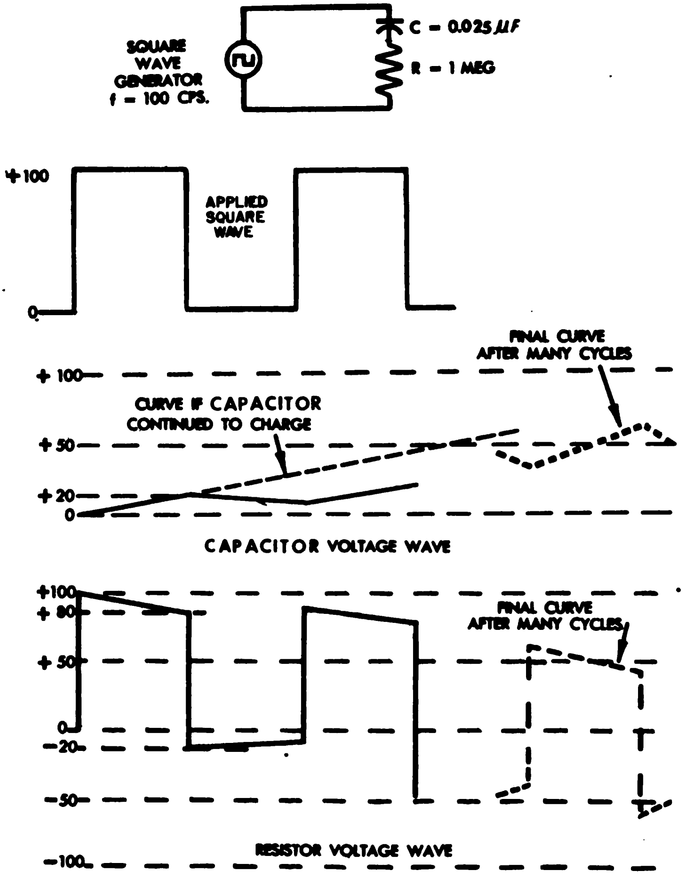


Figure 2–17.—Waveshapes for long time constant R–C circuit.

circuit is the capacitor voltage, which starts out at 20 volts and discharges 20 percent. As a result, the resistor voltage becomes minus 20 volts making all voltages around the circuit equal to zero and again fulfilling Kirchhoff's law.

The applied voltage showing waveshapes produced by long time constants (fig. 2–17) may either be positive voltages or voltages having a 50-volt d-c component plus a great number of a-c harmonics. In the latter case, the action of the capacitor is to charge to the d-c value after a number of cycles, this action being noticeable only during the first few cycles. Since a capacitor does not pass a d-c voltage, the resistor voltage appears as pure a.c. as soon as the capacitor is completely charged. You can see this from the resistor voltage wave in figure 2–17. Such a curve thus formed varies an equal amount on each side of the zero axis. Another important fact about this curve is that since each half cycle is a replica of the exponential curve, you can determine instantaneous values of voltages either by the exponential equations or the universal time constant chart. The chart was given earlier in this chapter.

Differentiation and Integration

An R–C voltage divider that is designed to distort the input voltage waveshape is known as a differentiator or integrator, depending on the locations of the output taps. The output from a differentiator is taken across the resistance while the output from an integrator is taken across the capacitor. Such circuits will change the shape of any complex alternating-voltage waveshape that is impressed on them, and this distortion is a function of the value of the time constant of the circuit as compared to the period of the waveshape. Neither a differentiator nor an integrator can change the shape of a pure sine wave. Both integrator and differentiator outputs are shown in figures 2–18, 2–19, and 2–21, but usually only one output is used in practical circuits.

SINE-WAVE INPUT.—If a 1,000-cycle sine-wave voltage that has a peak value of 100 volts is applied to an R–C voltage divider with an intermediate or short time constant, the sine wave will be shifted in phase. Figure 2–18 illustrates the case in which an intermediate R–C circuit is employed.

R is 10,000 ohms and *C* is 0.1 microfarad, so that the time constant of the circuit is $R \times C$ or 1,000 microseconds. The period of the 1,000 cycle voltage is $\frac{1}{1,000}$ seconds, or 1,000 microseconds. The time constant of the

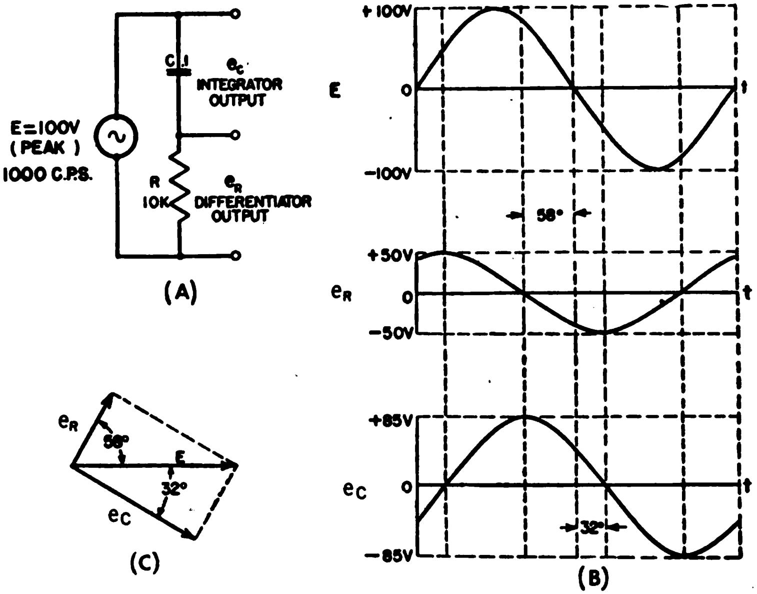


Figure 2–18.—R–C differentiator and integrator action on a sine voltage.

circuit is equal to the period of the voltage and the output will be distorted in phase and reduced in amplitude.

A-c voltage divider equations apply to this circuit, and the peak voltage across the resistor is found to be 50 volts, and across the capacitor 85 volts. Since these are vector quantities, they must be added vectorially to obtain the 100-volt input that Kirchhoff's laws require (fig. 2–18 (C)). This means that phase differences between the input, differentiator output, and integrator output will exist. Both outputs are sine waves as illustrated in figure 2–18 (B). In the case shown, the differentiator output e_R is a sine wave which leads the input voltage by about 58 degrees while the integrator output e_C is a sine wave which lags the input voltage by about 32 degrees. It may be shown, however, that the sum of the two outputs at every instant equals the instantaneous input voltage.

SQUARE-WAVE INPUT.—If the same circuit (fig. 2–18) has impressed on it a 1,000-cycle square-wave voltage

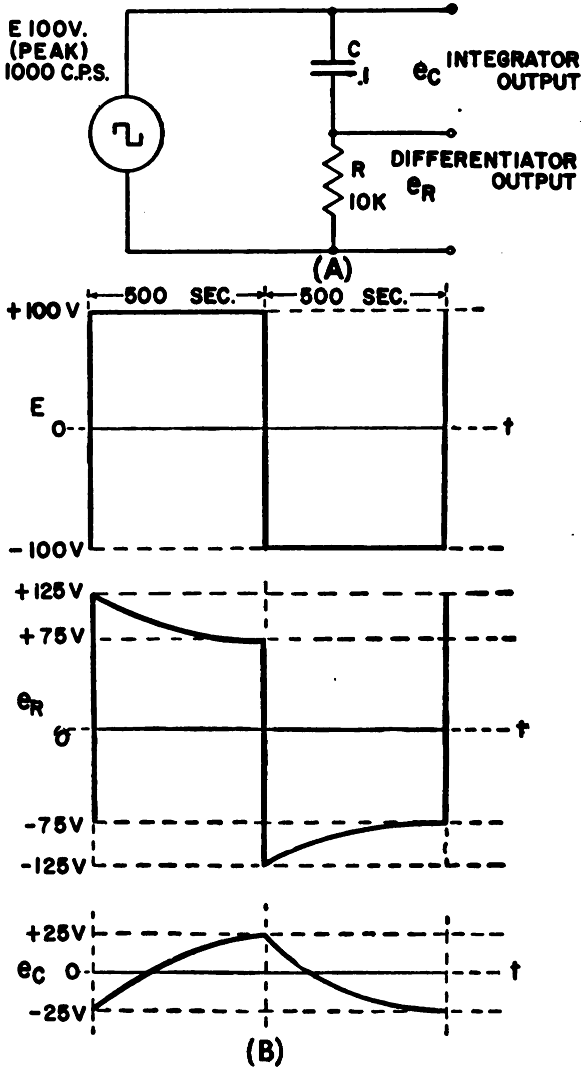
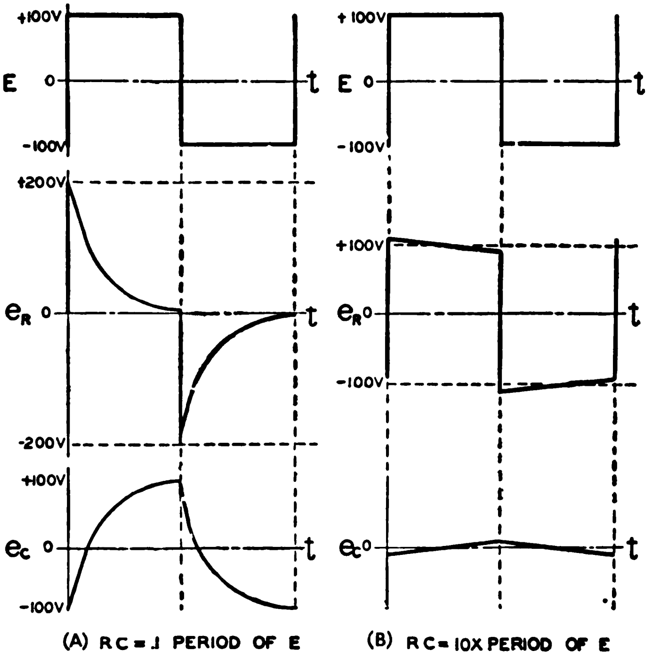


Figure 2–19.—R–C differentiator and integrator action on a square wave.

instead of a sine wave, the outputs shown in figure 2–19 (B) are obtained.

The square-wave voltage of 200 volts peak-to-peak is placed across the input of the circuit and the capacitor alternately charges positively and negatively by an amount determined by the R–C time constant. Since in this case the time constant is equal to the period of the square wave, the capacitor never fully charges during either half cycle and as a result the integrator output has a smaller amplitude than the input. Figure 2–19 (B) illustrates the outputs of the voltage divider. The two outputs must add at all times to the input voltage; consequently the differentiator output must be as shown. This output has a maximum amplitude that is greater than the input amplitude since the voltage left on the capacitor from the previous half cycle will add with the input voltage. The sum of these two voltages will appear as a voltage drop across the resistor at the instant the polarity of the input voltage changes.

Figure 2–20 shows the effect of different values of time constant on the output of the voltage divider. Figure



(A) RC = .1 PERIOD OF E (B) RC = 10X PERIOD OF E

Figure 2–20.—Effect of time constant on R–C differentiators and integrators.

2–20 (A) has a time constant equal to one-tenth the period of the input wave. The capacitor has time to become fully charged during a half cycle for all practical purposes. Such a circuit when used as a differentiator is often known as a PEAKER. The figure also makes apparent the fact that the peaks produced have an amplitude exactly twice the input amplitude.

Figure 2–20 (B) shows the effect of a time constant that is 10 times as long as the period. The differentiator output is almost the same as the input, and the integrator output has a very low amplitude. It is noticeable that the integrator output is a back-to-back sawtooth of very low amplitude. After amplification this sort of increase or decrease of voltage is a valuable part of some waveshapes used in electronics.

SAWTOOTH-WAVE INPUT.—If a back-to-back sawtooth voltage is applied to an R–C circuit with a time constant one-sixth the period of the input voltage, the result is as shown in figure 2–21.

The capacitor voltage e_C will follow the input voltage very closely if the time constant is short. The shorter the time constant, the closer it will follow. The result is that the integrator output very closely resembles the input. The amplitude will be slightly reduced and there will be a slight phase lag.

Since the voltage across the capacitor is increasing at a constant rate, the charging and discharging current will be constant. This means the output voltage of the differentiator will be constant during each half of the sawtooth input. The resulting square wave is shown in figure 2–21 (B). It can be seen that the integrator output adds to the differentiator output at every instant to equal the instantaneous input voltage.

MISCELLANEOUS INPUTS.—Various voltage waveforms other than those represented above may be applied to short R–C circuits for the purpose of producing, across the resistor, an output voltage with an amplitude proportional to the rate of change of the input. The shorter

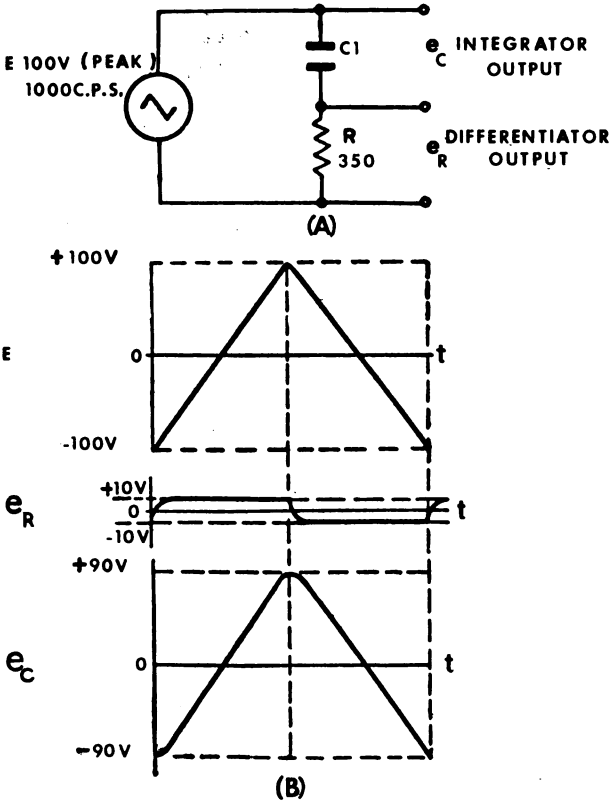


Figure 2–21.—R–C differentiator and integrator action on a sawtooth wave.

the R–C time constant is made with respect to the period of the input wave, the more nearly the voltage across the capacitor conforms to the input. Thus the differentiator output becomes of particular importance in very short R–C circuits.

The differentiator outputs for various types of input waves are shown in figure 2–22. In (A) the slope of the trailing edge of the sawtooth voltage is greater than that of the leading edge; therefore, the amplitude of the output voltage e_R is greater during this portion of the cycle. This same relationship of amplitude of the output

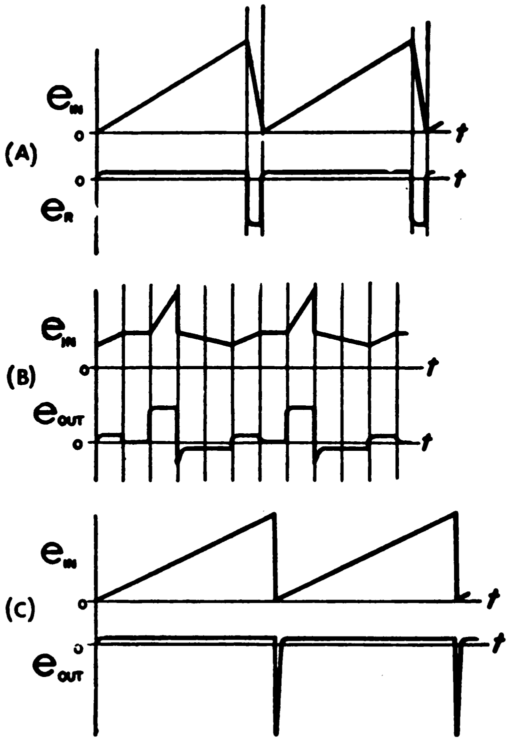


Figure 2–22.—Differentiator outputs of short R–C circuits for various input voltage waveshapes.

voltage e_R and the steepness of the input voltage rise or decay is illustrated in both (B) and (C).

TRANSIENT VOLTAGES IN L–R CIRCUITS

The operation of L–R circuits is similar to that of the R–C circuits just discussed since the inductor current in an inductor-resistor circuit rises exactly like capacitor voltage in an R–C circuit. With slight modifications, all the waveshapes obtainable in an R–C circuit are also obtainable in an L–R circuit.

L–R Charging Circuits

Figure 2–23 shows typical current and voltage curves produced in L–R circuits. In the L–R circuit shown, inductance L is assumed to be perfect; that is, it offers no resistance to d.c. When the switch is open, there is no voltage across the circuit, and because of this the current and voltage are zero. When you close the switch, the battery voltage E_{bb} is applied across the resistor R and inductance L. Current attempts to flow, but the inductor opposes this current by building up a back e.m.f., that, at the initial instant, exactly equals the input voltage E_{bb}. Since current cannot flow under this condition, there is no voltage across the resistor R. The two lower curves thus show that at the instant the switch is closed in the circuit, all the voltage is impressed across L and no voltage across R.

As current starts to flow, the voltage e_R appears across R, and the voltage e_L across L is reduced by the amount of e_R at that instant. A reduced voltage across the inductor means a less rapid increase in i_L and thus a less rapid increase in the resistor voltage. The e_L curve shows that e_L finally becomes zero when the current stops increasing while the e_R builds up gradually to the input voltage as the charging current rises. Under a steady state condition, the resistor is the only factor which limits the magnitude of the current.

Suppose you examine the three curves closely. Notice in the i_L (instantaneous current) curve that the current starts from zero and increases rapidly at first and then tapers off to practically a zero rate of increase at the end of 5 time constants. According to Ohm's law, the instantaneous values of the voltage across the resistor are directly proportional to the current flowing through it. Therefore, the e_R curve is similar to the i_L curve.

Since the sum of voltage around a series circuit (the circuit illustrated is a series L–R circuit) equals zero, and since the applied voltage is constant, the inductor voltage e_L is always equal to the difference between E_{bb} and e_R. The inductor voltage equals E_{bb} when the resistor e_R equals zero, and equals zero when e_R is at maximum value. If you reverse the polarity of the e_L curve, you

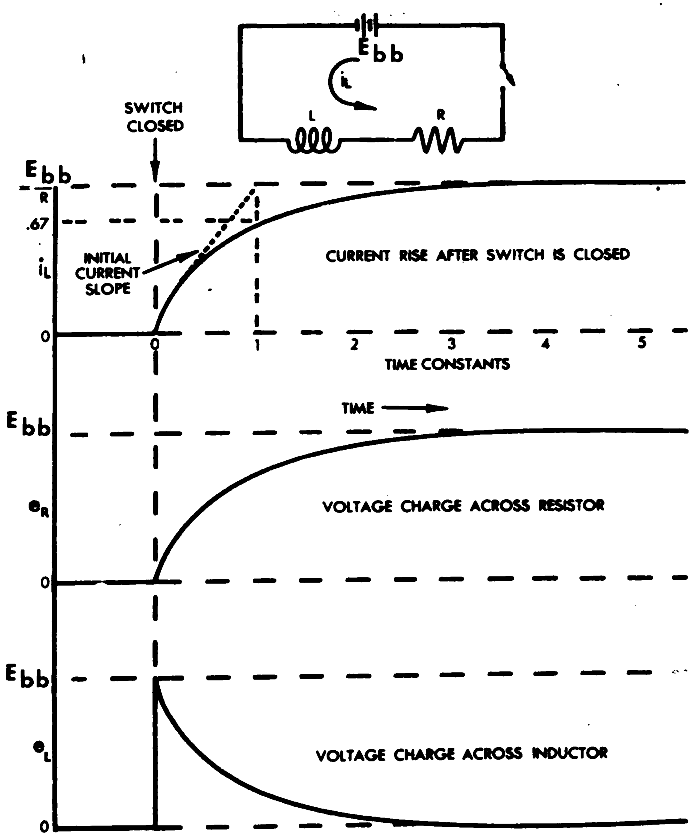

Ebb
iL
L
R
SWITCH CLOSED
Ebb/R
.67
iL
INITIAL CURRENT SLOPE
CURRENT RISE AFTER SWITCH IS CLOSED
0
0
1
2
3
4
5
TIME CONSTANTS
TIME
Ebb
eR
VOLTAGE CHARGE ACROSS RESISTOR
0
Ebb
eL
VOLTAGE CHARGE ACROSS INDUCTOR
0

Figure 2–23.—Current and voltage curves in L–R charging circuit.

may also consider it is representing the back e.m.f. in the inductor. (The polarity of back e.m.f. is always opposite that of the applied voltage.) At first, the back e.m.f. is high since the current tends to jump to a value of $\frac{E_{bb}}{R}$. This large back e.m.f. opposes the applied voltage and keeps the current small. In order for the back e.m.f. to be sustained at its initial high value, the current must continue to increase at the initial rate (dotted line in i_L curve). This is not possible, however, for as the current increases, the resistor voltage increases proportionately. The increase in resistor voltage effectively reduces the inductor voltage. Hence, the back e.m.f. decreases while current and resistor voltage increase. All three curves change along exponential curves as shown in figure 2–23.

L–R Time Constant

The meaning of the time constant in an L–R circuit is similar to the meaning of the time constant in an R–C circuit. An L–R time constant is defined as the time required for the current through an inductor to increase to 63.2 percent of its maximum value. The formula for the L–R time constant is

$$t.c.=\frac{L}{R}$$

where $t.c.$ represents the time constant in seconds, L the inductance in henrys, and R the resistance in ohms.

The ratio $\frac{L}{R}$ also represents the time required for the resistor voltage to equal 63.2 percent of the applied voltage, and the time required for the inductor voltage to equal 36.8 percent of the applied voltage.

L–R Charging Problems

The shapes of the curves for current and voltage in the L–R circuits are like those in the R–C circuit. Therefore, you can use the rough approximation curves and the universal time constant curves for determining instantaneous values of current and voltage.

Study the following sample problems to learn how to find instantaneous values of current and voltage in L–R circuits. These problems are solved by the curve chart method.

Examples:

Problem 1.—Using the values indicated in the circuit of figure 2–24 find the amount of current 200 microseconds after the switch is closed.

Solution:

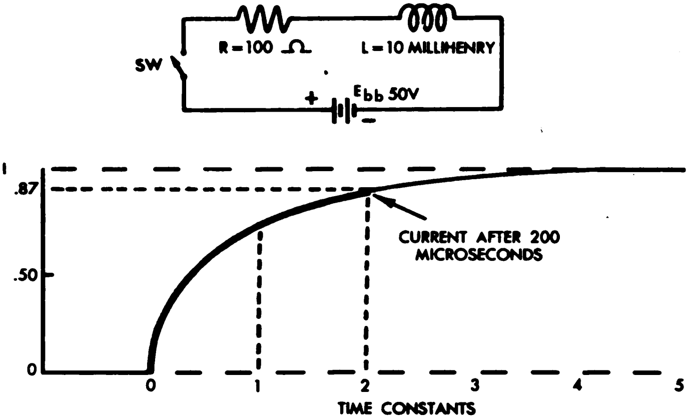


Figure 2–24.—Solving L–R problems.

Find the time constant by the formula

$$t.c.=\frac{L}{R}$$

$$t.c.=\frac{0.01 \text{ henry}}{100 \text{ ohms}}$$

$$t.c.=10^{-2}\times10^{-2}$$

$$t.c.=10^{-4}$$

$$t.c.=100 \text{ microseconds.}$$

Since the time constant is 100 microseconds, and the time involved is 200 microseconds, the charging action will last 2 time constants. From the rough approximation curves in figure 2–9 or the universal time constant chart in figure 2–10, you find that the current in the inductor is 87 percent of maximum at 2 time constants. Calculating the maximum current by the formula

$$I=\frac{E_R}{R}$$

$$I=\frac{50}{100}=0.5 \text{ amperes.}$$

Since the maximum current is 0.5 amperes, it will be

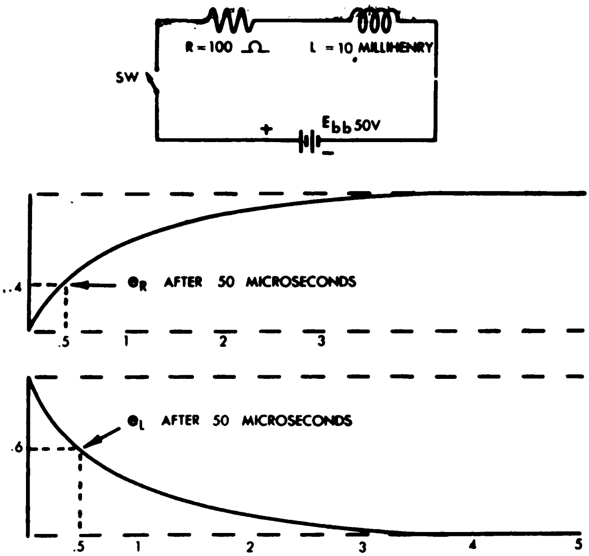


Figure 2–25.—Solving L–R problems.

0.87 of 0.5 or 0.435 amperes at the end of 200 microseconds.

Problem 2.—Using the same circuit repeated in figure 2–25 and the same time constant find e_R and e_L 50 microseconds after the switch is closed.

Solution:

The rough approximation curves do not contain a 50 microsecond time. Therefore, refer to the universal time constant chart. (Curve *A* is the resistor voltage curve; curve *B* the inductor voltage during charge.) From curves *A* and *B*,

$$e_R = 0.4 \text{ of maximum voltage}$$

$$e_L = 0.6 \text{ of maximum voltage}$$

Therefore,

$$e_R = 0.4 \times 50 \text{ volts} = 20 \text{ volts}$$

$$e_L = 0.6 \times 50 \text{ volts} = 30 \text{ volts}$$

L–R Discharging Problems

The curves shown in figure 2–26 are formed when the source of energy is removed from a series L–R circuit, and the energy stored in the magnetic field is returned to the circuit by the collapse of the field. At the initial instant when you open the switch, the resistor voltage e_R attempts to decrease to zero, but the inductance of the inductor resists any change in current and tends to maintain the current flow. This action of the inductor occurs because opening the switch removes the applied voltage, causes the current flow in the circuit to tend to cease, and starts the collapse of the magnetic field about the inductor. This collapse induces a voltage in the inductor, which causes the current to flow in the same direction that it did when the circuit was charged. Now the self-induced voltage is the source voltage in the circuit and its polarity is opposite. As the discharge

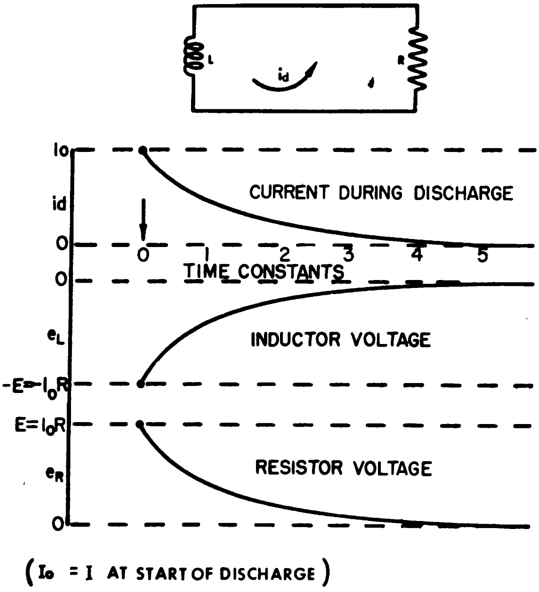


Figure 2–26.—Voltage and current curves during collapse of inductor field.

current i_d begins to decrease, the voltage across R decreases proportionately.

Since $e_R + e_L = 0$, as previously mentioned in L–R charging, the value of e_L also decreases. This change continues as shown in the e_R and e_L curves until the e_R and e_L voltages are both zero and no current is flowing. The discharge current at the end of one time constant is 36.8 percent of maximum.

Study the following sample problems to learn how to find the instantaneous values of current and voltage in L–R discharging circuits. The methods employed are like those used previously in R–C discharge circuits. The formula for the L–R time constant previously discussed applies here too.

Solving by use of curves.

Problem 1.—Find the current 150 microseconds after the current has started to decrease in the circuit of figure 2–27.

Solution:

Finding the time constant, $t.c. = \frac{L}{R} = \frac{0.01}{100} = 10^{-4} = 100$ microseconds. Since the time involved is 150 microseconds, the number of time constants is 1.5. Then, since the current starts high and decreases to zero (refer to curve B in the universal time constant chart), this curve shows that the current drops to 22 percent of its initial value after 1.5 time constants. The initial current is 100 ma. Therefore, the current after 150 microseconds is equal to 100×0.22, or 22 ma.

Problem 2.—Using the same circuit, find the resistor and inductor voltages 250 microseconds after the current starts decreasing.

Solution:

The time given, 250 microseconds, constitutes 2.5 time constants. The universal time constant chart shows that after 2.5 time constants, the voltage drops to 8 percent of its initial value. Therefore, since the maximum voltage $E_R = IR = 0.1$ ma. $\times 100$ ohms, or 10 volts,

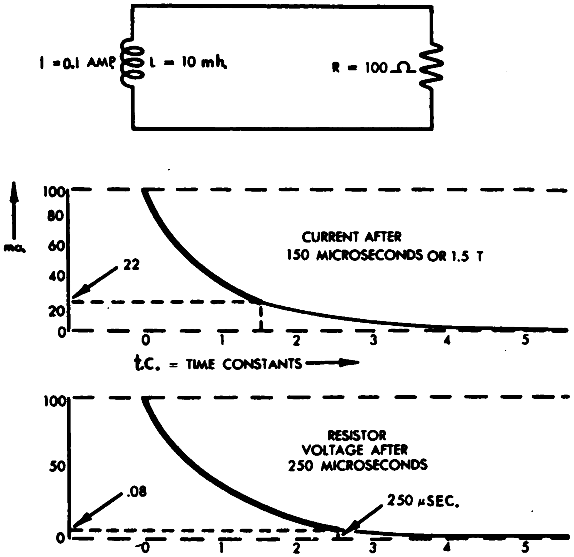


Figure 2–27.—Solving L–R problems.

the voltage e_R, after 2.5 time constants, equals 10×0.08, or 0.8 volt.

In finding the inductor voltage, utilize the fact that the inductor voltage is the source voltage in an L–R discharging circuit and, at all time, is equal to the voltage across the resistor. Therefore, since the resistor voltage, e_R, is 0.8 volt after 2.5 time constants, the inductor voltage, e_L, equals 0.8 volt.

Effect of Varying *R* and *L*

As shown in figure 2–28, varying the size of *R* and *L* in an L–R circuit changes the charge and discharge curves. The changes result from the fact that any change of either *R* or *L* changes the L–R time constant. The solid-line curves shown represent the instantaneous values of current and voltage during charge and discharge when the time constant is 100 microseconds.

Decreasing the time constant sharpens the slope of both the charge and discharge curves. Notice, for example, the dashed-line curve. It represents the effect of

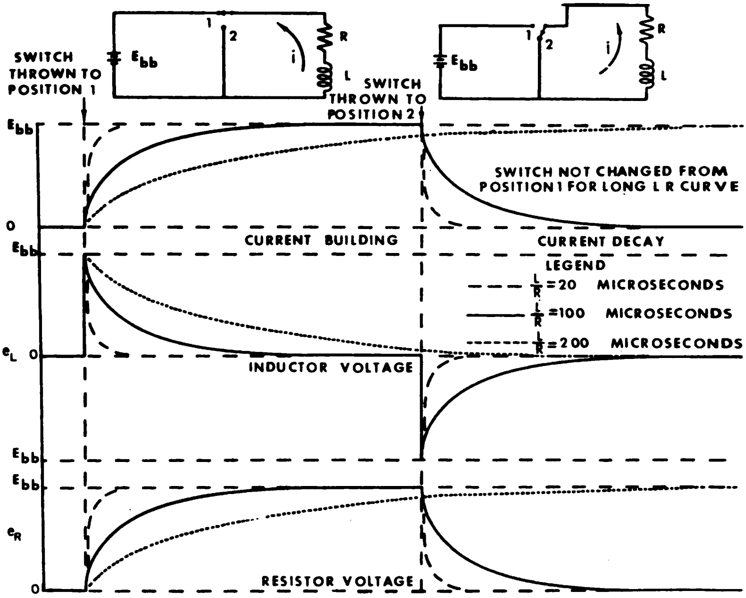


Figure 2–28.—Effect of varying *R* and *L*.

reducing the time constant in the circuit to 20 microseconds. A time constant may be reduced either by decreasing *L* or by increasing *R*. Decreasing *L* decreases the opposition to current flow change and permits current to build up and decay more quickly. Increasing *R* does not affect the inductance, but it does reduce the maximum current flow. Since the maximum value is lower, the current reaches its maximum value more quickly. Because the current builds up very rapidly in a circuit with a short time constant, the voltage across the resistor (e_R), which keeps in step with the current, similarly builds up rapidly. The inductor voltage (e_L), which drops just as fast as the resistor voltage increases, likewise forms a sharp peak. When current is decaying, the resistor voltage and the inductor voltage similarly decay much more rapidly than in a circuit with a longer time constant.

Increasing a time constant produces curves with gradual slopes. For example, notice the dotted-line curve. It represents the curve formed by a 200-microsecond time constant. The rise of current is slower than before because the inductor furnishes more opposition to the flow of current. The effect is noticeable also in the other 200-microsecond curves, where the fall of voltage across the inductor is very slow, and the rise of voltage across the resistor requires appreciably more time than with a 100-microsecond time constant. Similarly, the decay of current takes longer since a large inductor stores more energy than a small one, and consumes more time in dissipating it through the resistor. If the time constant were lengthened by decreasing *R*, the same relationships would hold true, because the smaller resistor would dissipate the energy stored in the inductor more slowly. You can see why by analyzing the basic concepts; energy dissipated is power, and power is I^2R. Therefore, if the current (I) is constant and the resistance (R) is decreased, the power dissipation is less at any single instant. This means that if the rate of dissipation is decreased by using a smaller resistor, a longer time is required to dissipate the energy stored in the inductor.

LONG L–R TIME CONSTANT.—The waveshapes illustrated in figure 2–29 are those produced when a square wave is applied to an L–R circuit with a long time constant. The time constant used is extremely long, being ten times the period. One important fact to observe here, other than the waveshapes, is how the inductor voltages change polarities during the time that the inductor acts as the source of voltage in the circuit. During current buildup, the end of the inductor that current enters may be called the negative end, for during this time the inductor acts just like any resistance or, for that matter, like any other load. However, when the source of voltage is removed, the inductor becomes a source, and like a battery, the current leaves the negative end and enters the positive end. Since current flows in the same direc-

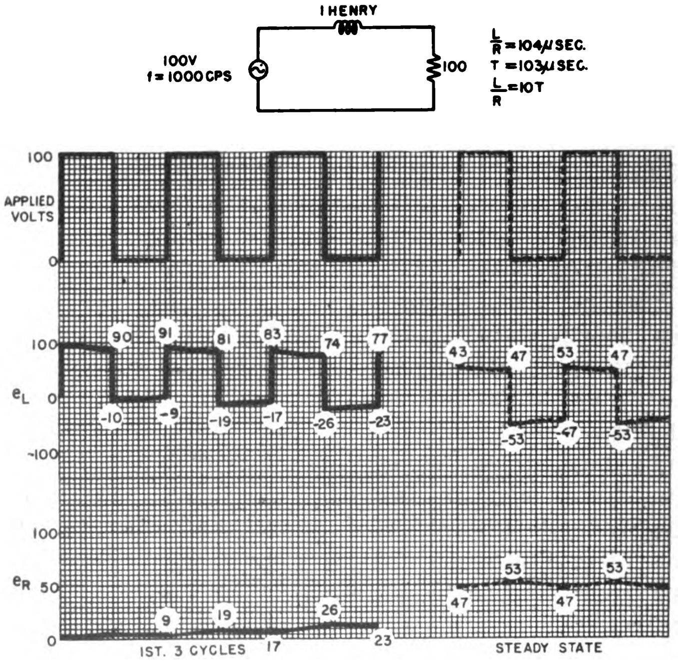


Figure 2–29.—Effect of long L–R time constant.

tion during either buildup or decay, the polarity of voltage actually reverses at each half cycle.

SHORT L–R TIME CONSTANT.—The waveshapes illustrated in figure 2–30 are due largely to using a smaller inductor. The small inductor allows the current to reach a maximum value very quickly and thus causes the resistor voltage to be nearly a replica of the input voltage. Another result is that the inductor voltage is differentiated into a sharp peaked wave.

MEDIUM L–R TIME CONSTANT.—The waveshapes produced by applying a square wave to a medium L–R time constant circuit with the values indicated are shown in figure 2–31. The time constant employed is equal to the time duration of one half cycle of the input square wave. Therefore, the current reaches 0.63 amperes, or 63 percent of its maximum charge value during the first half

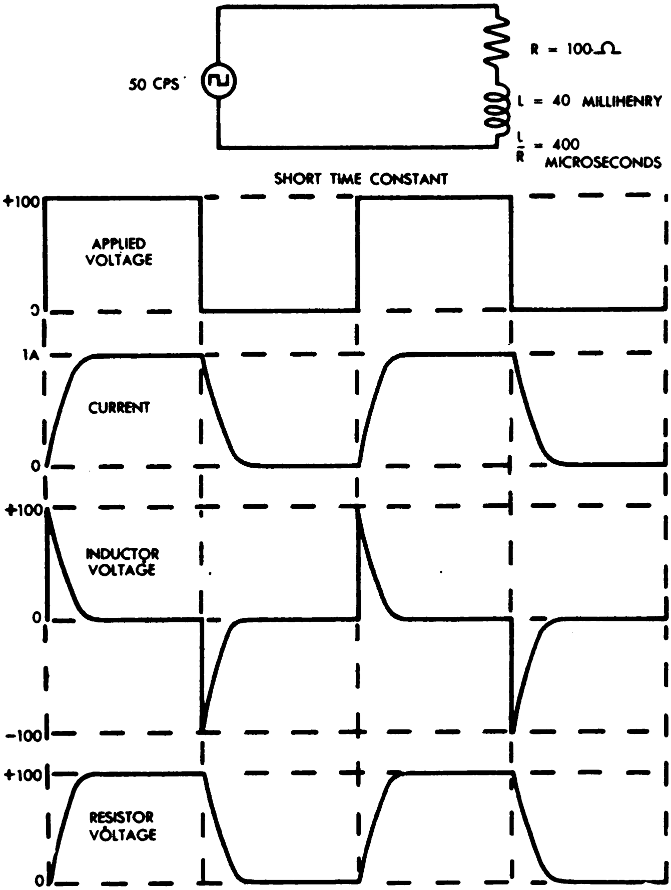


Figure 2–30.—Effect of short L–R time constant.

cycle. During the second half cycle, when the applied voltage is zero and the inductor furnishes the current, the current drops to 0.23 amperes or 37 percent of 0.63 amperes. Since the applied voltage varies between 0 and 100 volts, there exists a 50-volt d-c component in the waveshape. This d-c component causes the current during charge and discharge to vary about an average value of 50 volts divided by 100 ohms, or 0.5 amperes. However, because of the inductance in the circuit, the current

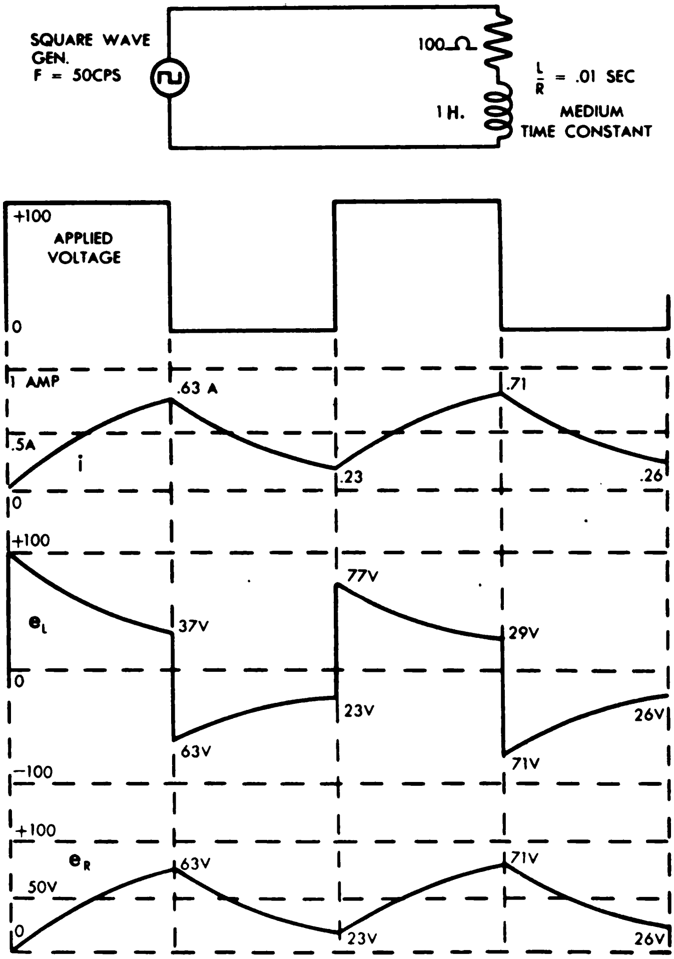


Figure 2–31.—Effect of medium L–R time constant.

does not reach this average value immediately but approaches it gradually. It does this by starting each current half cycle a little higher. For example, the average value in the second half cycle is 0.42 amperes; the current range is 0.63 amperes to 0.23 amperes. In the third half cycle it is 0.47, the range being 0.23 to 0.71, and in the fourth, the average is about 0.49 with a range of 0.71 to 0.26 amperes. During the first positive half cycle the inductor voltages are entirely positive.

With succeeding cycles of the input voltage, these voltages work down until they center around the zero voltage line; that is, about as much of the inductor voltage is above zero as is below. The e_R voltage wave increases in value with each half cycle of the applied voltage thus indicating that a d-c voltage component exists in the square wave. Notice that at the end of the first half cycle the resistor voltage is 63 volts and at the beginning of the second half cycle it is 71 volts. After 5 half cycles of the input voltage, the waveshape repeats at the values in the last half cycle shown.

Frequently, the applied voltage does not contain a d-c component at all, but is a pure square wave. In that case waves like those shown in figure 2–32 are produced. The current and voltage waves produced when the pure square

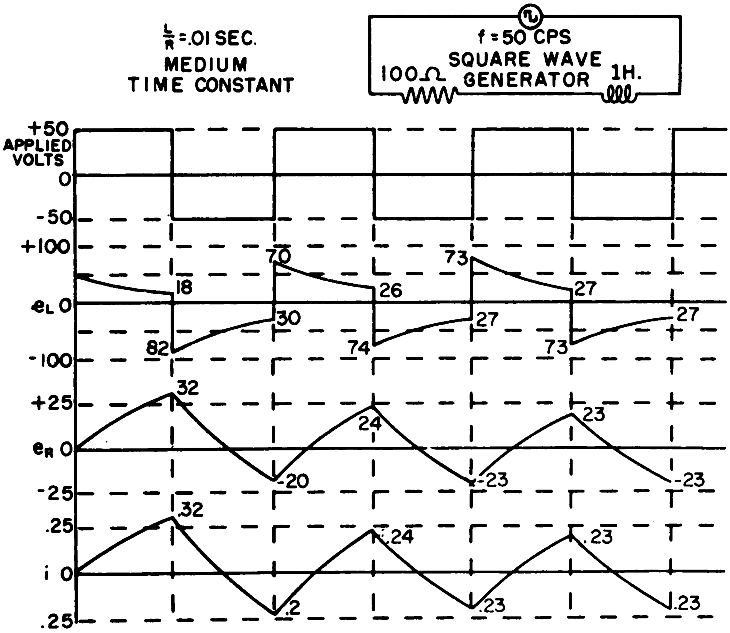


Figure 2–32.—L–R waves when input contains no d-c component.

wave is applied at the L–R circuit shown are identical to those produced when the square wave has a d-c component, but the resistor voltages are 50 volts less than their original value.

Values of Current and Voltage

An easy means of determining the voltage or current any time a square wave is applied to an L–R circuit is to reduce the circuit to a simple d-c circuit. For example, the L–R circuit in figure 2–33 can be represented by two equivalent circuits—one (t_1) representing the 0.01 second of the square wave when 100 volts is applied, and the other (t_2) the 0.01 second when zero voltage is applied.

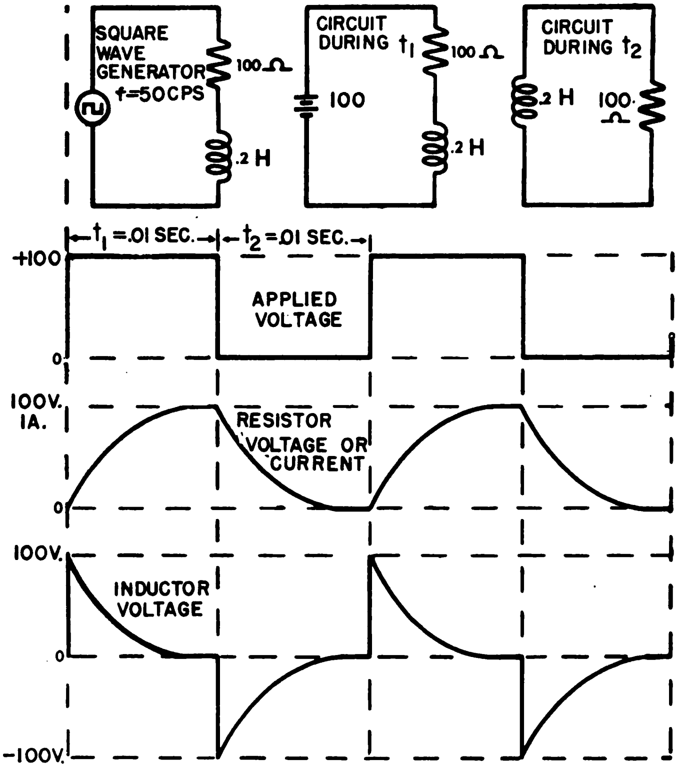


Figure 2–33.—Equivalence in L–R circuit.

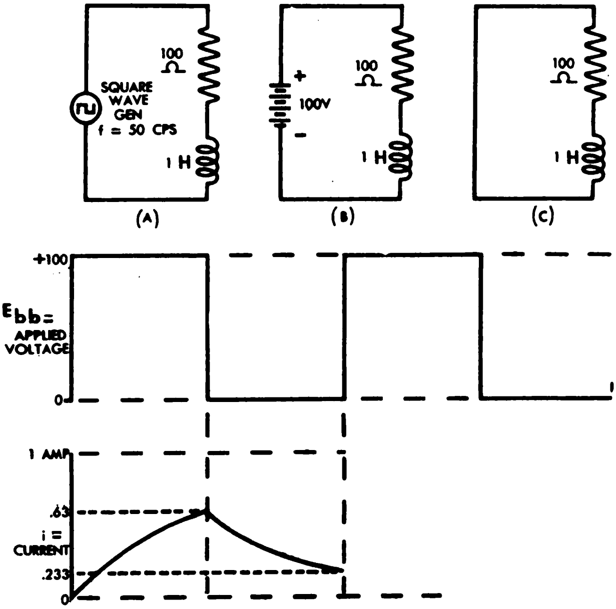


Figure 2–34.—Circuit with voltage and current curves.

During t_1 the circuit may be reduced to an L–R circuit in which the generator is replaced by a battery with a voltage equal to the generator voltage. This is proper since the current rises from zero when a battery is used, just as it does when the square wave voltage is applied. Therefore, if you know the time the voltage is applied and the values of L and R, you can compute the value of current at any time.

During time t_2 you can reduce the L–R circuits shown to a circuit containing a resistor and an inductor in which current is flowing.

To understand how to apply these equivalent circuits in finding either current or voltage at a fixed time, study the following example:

Problem.—In the circuit in figure 2–34 find the current at the end of each of the first two time constants.

Solution:

When the square wave generator is producing the positive half cycle, it applies a constant 100 volts to the circuit in (A) of figure 2–34. The equivalent circuit is shown in (B) of figure 2–34. During this time current rises toward its maximum $\frac{E}{R}$ value of 1 ampere. Since the time constant is 0.01 second, and since the time of a half cycle is also 0.01 second, the 100 volts is applied for one time constant. Therefore, at the end of one time constant, the current equals 63 percent of maximum, or 0.63 amperes.

The circuit condition when the applied voltage is zero is shown in (C) of figure 2–34. Initially the current is 0.63 ampere, as just determined. The time of this half cycle is also 0.01 second, and likewise one time constant. Since current in an L–R circuit drops to 37 percent of its initial value in one time constant, it here equals 37 percent of 0.63 ampere or 0.233 ampere at the end of the second time constant.

TRIANGULAR WAVES IN R–C AND L–R CIRCUITS

All the voltages thus far described in connection with R–C and L–R circuits were voltages of sudden changes, such as a voltage which changes to full magnitude when a switch is closed or as the voltage of the steep wavefront of a square wave. However, not all voltages applied to R–C and L–R circuits are sudden-change voltages. For example, triangular or sawtooth waveshape voltages, such as those shown in (A) of figure 2–35, are commonly used in electronic circuits. It is important that you know the effect that these waveshapes have on long and short time constant R–C and L–R circuits.

Response to Triangular Wave Input

For ease of understanding the effect of triangular waves in L–R and R–C circuits, first examine the first rise in voltage when this wave is applied to the R–C circuit. (See fig. 2–35 (B).) The voltage wave enlarged several times for easier study is shown in (C) of figure 2–35. Notice that the voltage starts at zero and keeps on increasing at a constant rate. The triangular waveshape is thus different than that of a square wave, in which the voltage rises quickly then levels off. Likewise, the effect of a triangular wave on the capacitor is different than that of a square wave. With a square wave, the capacitor charges exponentially to the value at which the square wave levels off, but with a triangular wave the voltage increases as the capacitor charges toward it (note the triangular shape of the wave), and the capacitor must continually charge toward a new higher value.

Initially when the voltage e_i is applied to the circuit (fig. 2–35 (B)), the current rises with the voltage because the capacitor is initially uncharged. This current charges the capacitor to a voltage that opposes the applied voltage. This should decrease the current, but since the input voltage is continually increasing, the opposing voltage is more than overcome by the applied voltage, and the current actually increases, but less rapidly than before. As the current increase continues, the capacitor becomes charged to a higher voltage and the rate of current increase falls off until finally the current is high enough to raise the capacitor voltage at the same rate at which the input voltage waveshape increases. At this point, the current becomes constant, and the capacitor voltage slope is essentially the same as the input voltage slope. At this time the constant current through the resistor causes a constant resistor voltage.

Since current increases from zero to a constant value in an R–C circuit, its rate of increase is exponential. You can use the curve for e_R as the current curve. Since both the current and voltage are exponential, the current and e_R at the end of one time constant will be 63.2 percent of the voltage applied at that time. After 5 time constants, the current will level off to a value equal to the voltage applied at the end of the first time constant; that is, at the value $e_{\max}$ shown in figure 2–35.

The maximum value of resistor voltage ($e_{\max}$) is a function of the time constant and the rate of rise. It is expressed by the equation

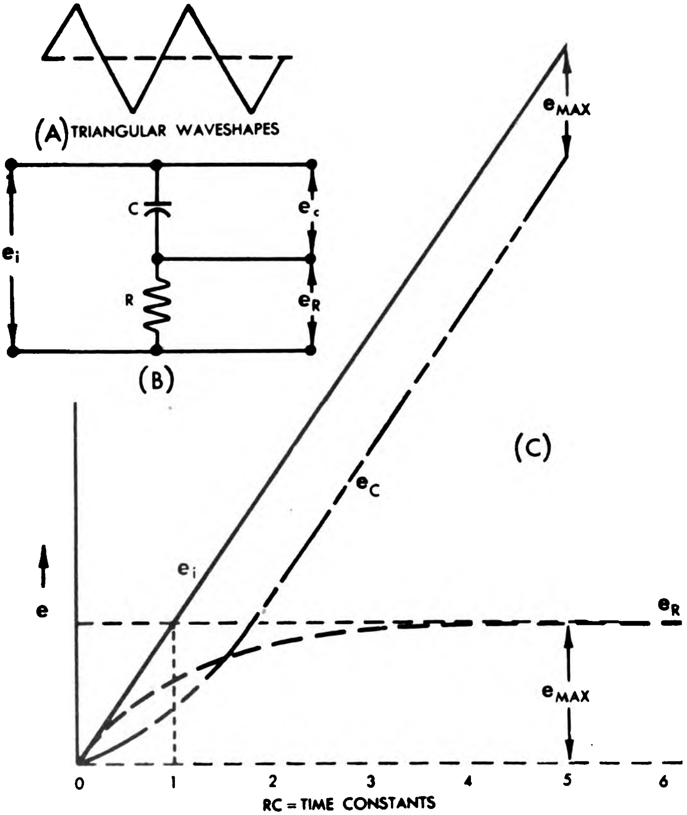


Figure 2–35.—Response of R–C circuit to triangular wave input.

$$e_{\max}=\frac{\Delta e_i}{\Delta t}RC$$

In this equation,

$e_{\max}$=Maximum resistor voltage

R=Resistance in ohms

C=Capacitance in farads

$\frac{\Delta e_i}{\Delta t}$=Rate of change of input voltage in volts per second. (The symbol Δ followed by a quantity indicates the amount of change in the quantity.)

Example:

Problem.—Assuming that the triangular waveshape illustrated has a frequency of 50 c.p.s. and is applied to the circuit in figure 2–36, find the maximum resistor voltage.

Solution:

Inspection shows that the change of the input voltage e_i is 100 volts during one half cycle. You can represent this change in input voltage as Δe_i. Since the frequency is 50 c.p.s., one cycle is $\frac{1}{50}$ or 0.02 second. Therefore, the half cycle is 0.01 second long. You can represent the change in time as Δt. Therefore, Δt, during the 100-volt rise, is 0.01 second.

Substituting in equation,

$$e_{\text{max}}=\frac{\Delta e_i}{\Delta t}RC$$

$$e_{\text{max}}=\frac{100\times 10^5\times 0.02\times 10^{-6}}{0.01}$$

$$e_{\text{max}}=\frac{10^2\times 10^5\times 2\times 10^{-2}\times 10^{-6}}{10^{-2}}$$

$$e_{\text{max}}=2\times 10$$

$$e_{\text{max}}=20 \text{ volts.}$$

The capacitor continues to charge at the same rate until the rise of input voltage ceases. As the voltage abruptly starts decreasing during the second half of the cycle, the charging rate of the capacitor decreases until the voltage drops to the same value as the capacitor voltage. At this point, the current is zero. As the applied voltage continues to decrease, it falls below the capacitor voltage, and the capacitor starts discharging. This reverses the current, causing a negative voltage across the resistor. Again, the changeover cannot occur any faster than the capacitor can discharge, so the current, and consequently the resistor voltage change is exponential. The decrease

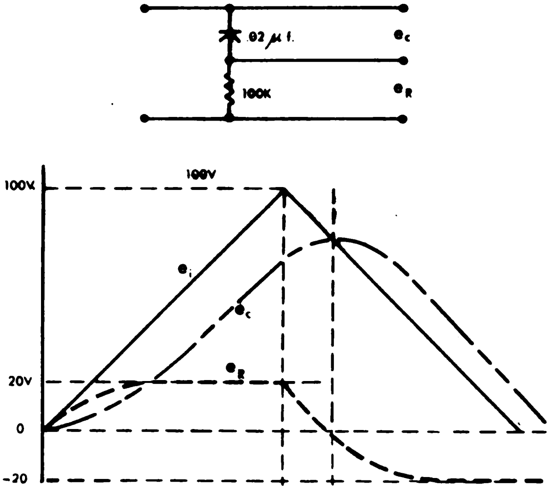


Figure 2–36.—Capacitor and resistor voltages with triangular wave input.

in voltage is at the same rate ($\frac{\Delta e_i}{\Delta t}$); so the negative voltage is also 20 volts.

Effect of Varying the Time Constant Ratio

The effect of varying the time constant is indicated in figure 2–37. The waveshape of the previous problem repeated over several cycles is shown in (A) of figure 2–37. As previously mentioned in connection with square waveshapes, the capacitor waveshape resembles the input when the R–C is short and resistor waveshape resembles it when the R–C is long. This is true for triangular waveshapes also. The R–C is twice the time of the applied voltage (fig. 2–37 (B)); so the resistor waveshape very closely follows the input. If the R–C is shortened to equal the time the voltage is increasing, the resistor voltage reaches 63 percent of the applied voltage after one time constant. However, if the R–C is shortened to one-fifth of the time the voltage is increasing, the capacitor charging rate quickly reaches the rate of applied voltage, after which the resistor voltage is constant. Notice that this

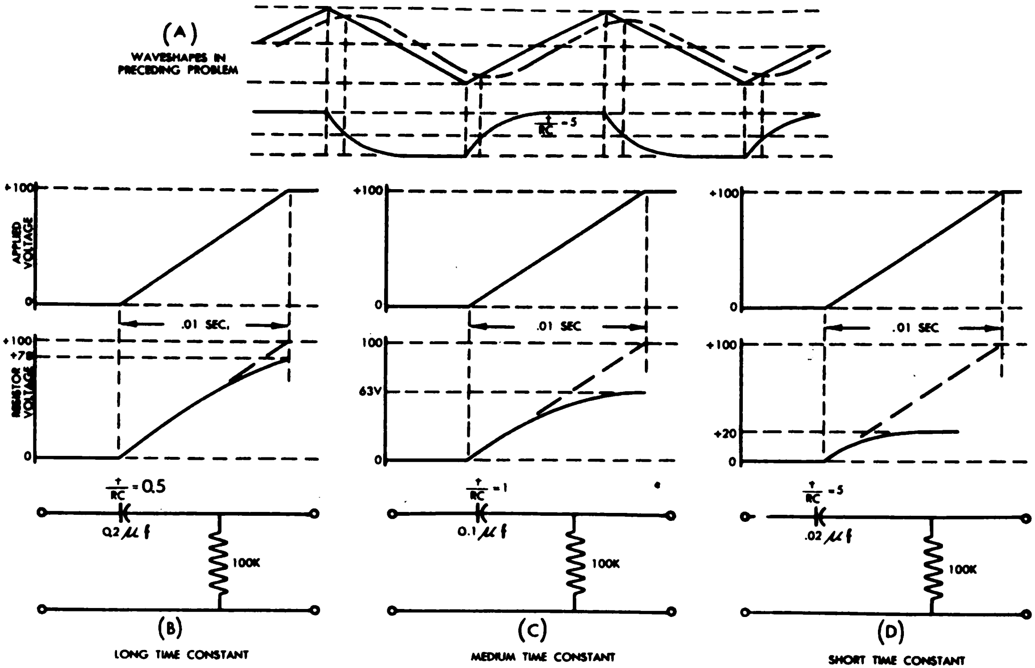


Figure 2–37.—Varying the time constant with triangular wave input.

condition causes the resistor voltage to approach a square wave (fig. 2–37 (A)). This is differentiation which was described earlier. When a triangular wave is integrated by use of a long time constant, the capacitor waveshapes will resemble a hyperbolic curve. (See fig. 2–37 (B).)

An L–R circuit has the same effect on triangular waveshapes as an R–C circuit provided you observe the following changes:

1. The waveshape which appears across the resistor in the R–C circuit appears across the inductor in the L–R circuit, while the wave which appears across the capacitor in the R–C arrangement appears across the resistor.
2. The time constant is $\frac{L}{R}$ instead of $R \times C$.
3. The term $\frac{\Delta e_i}{\Delta t}$ is used to obtain the final voltage across the inductor.

When a triangular voltage waveshape is applied to an L–R circuit, the current increases at a constant rate. This, in turn, causes the inductor voltage to become constant after 5 time constants. You can express the maximum inductor voltage by the equation

$$E_L = \frac{\Delta e_i}{\Delta t} \times \frac{L}{R}$$

where E_L is the maximum inductor voltage, $\frac{\Delta e_i}{\Delta t}$ the rate of change of voltage, L the inductance in henrys, and R the resistance in ohms.

PHASE SHIFTING

Review of Phase Relations in R–C and R–L Circuits

For an understanding of phase splitting and phase shifting the following current and voltage relationships in R–C and R–L circuits should be recalled:

1. The current and voltage are in phase in a resistor.
2. The current leads the voltage by 90 degrees in a capacitor.

3. The current lags the voltage by 90 degrees in an inductor.
4. When a resistor and capacitor are connected in series, the voltage across the resistor leads the voltage across the capacitor by 90 degrees.
5. When a resistor and an inductor are connected in series, the voltage across the resistor lags the voltage across the inductor by 90 degrees.

Phase-Splitting Circuits

When a resistor and a capacitor are connected in series with an applied sine-wave voltage (fig. 2–38), two output voltages can be obtained which differ in phase by 90 degrees. The voltage across the resistor *R* is always in phase with the current in *R*, while the voltage across capacitor *C* is always 90 degrees out of phase with the current flowing in *C*. Therefore, the voltage outputs across resistor *R* and capacitor *C* (E_R and E_C) are 90 degrees out of phase with each other. Such a circuit, in which two outputs are used, is called a PHASE-SPLITTING CIRCUIT.

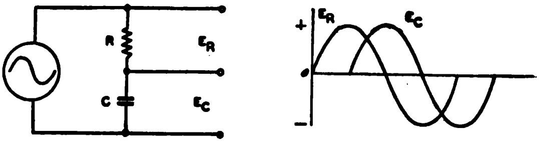


Figure 2–38.—R–C phase-splitting circuit.

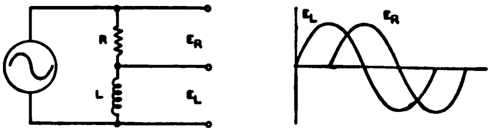


Figure 2–39.—R–L phase-splitting circuit.

A similar condition results when a resistor and an inductor are used (fig. 2–39), except that the direction of the phase shift is reversed. Since an inductor in an a-c circuit causes the current to lag the voltage by 90 degrees, the output voltages of the phase-splitting circuit differ by 90 degrees, and the voltage across the resistor lags the voltage across the inductor.

R–C Phase-Shifting Circuit

The R–C phase-splitting circuit (fig. 2–38) produces a phase shift or phase difference of 90 degrees between the voltage across the resistor and the voltage across the capacitor. The shift in phase between one of these voltages and the supply voltage depends upon the ratio between the resistance and the reactance in the circuit.

When the resistance is large with respect to the capacitive reactance (fig. 2–40 (A)), most of the opposition to current flow is caused by the resistance, and the effect of the capacitor is small. The current through the circuit is therefore almost in phase with the supply voltage. Since the voltage drop across the resistor is in phase with the current, E_R is almost in phase with E_O. The output voltage across the capacitor is nearly 90 degrees out of phase with the voltage across the resistor. This phase relationship is shown in figure 2–40 (B) and in the vector diagram (C).

When the resistance is small with respect to the capacitive reactance (fig. 2–40 (D)), most of the opposition to current flow is caused by the capacitor. The current in the circuit therefore leads the applied voltage nearly 90 degrees. The voltage across the resistor, which is in phase with the current, is approximately 90 degrees out of phase with the applied voltage. Therefore, the voltage across the capacitor, which is 90 degrees out of phase with the voltage across the resistor, is nearly in phase with the supply voltage. This phase relationship is shown in figure 2–40 (E) and the vector diagram (F).

A phase-shifting circuit, therefore, may be composed of a capacitor and a variable resistor connected across a sine-wave supply voltage. The phase of the output voltage is shifted in relation to the input voltage by variations in the amount of resistance. When the output voltage is across the capacitor, very little resistance in the circuits gives an output almost in phase with the

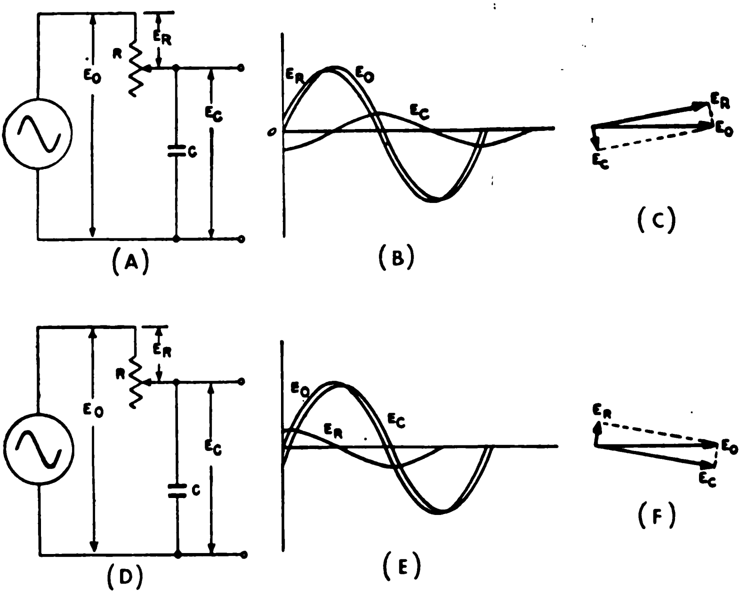


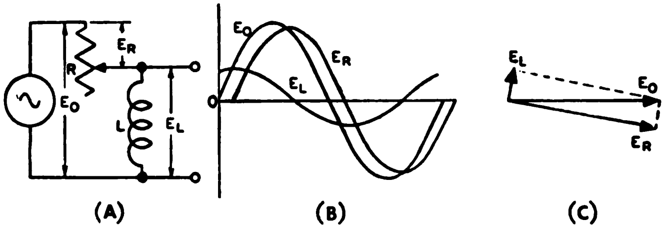
Figure 2–40.—R–C phase-shifting network.

input. A high resistance, on the other hand, gives an output voltage which lags the input voltage.

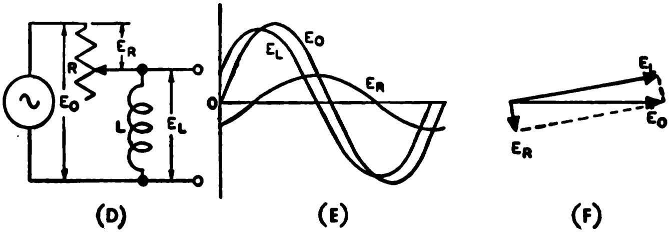
The output can also be taken across the resistor. In this case the direction of phase shift is reversed. Increased resistance brings the output and input voltages more nearly into phase with each other, while decreased resistance results in a greater degree of phase shift.

R–L Phase-Shifting Circuit

Substitution of an inductor for the capacitor in the circuit of figure 2–40 produces an R–L phase-shifting circuit (fig. 2–41 (A) and (D)). The phase relations between the input and output voltages can be seen in figure 2–41 (B) and (E) and in the vector diagrams (C) and (F). Since in the case of inductance the current lags the voltage, the voltage across the resistance always lags the voltage across the inductance. The output voltage normally is taken from across the inductance. An increase in the resistance of the phase-shifting circuit



R IS GREATER THAN X_L



X_L IS GREATER THAN R

Figure 2–41.—R–L phase-shifting network.

increases the difference in phase between input and output voltages. A decrease in the resistance brings the input and output voltages more nearly into phase with each other.

One disadvantage of the R–C and R–L types of phase-shifting circuits is that the voltage across either the capacitor or inductor becomes smaller and smaller as the resistance is increased. As a result, these networks are not always satisfactory in a circuit in which a definite output voltage is required. They are practical, however, as networks which can be adjusted to give a definite fixed value of phase shift.

Use of Phase-Shifting Circuits

Phase-shifting circuits are used for two general purposes:

1. To obtain a definite degree of phase shift which may be fixed or adjustable within small limits. For example, this type of circuit is used to correct a shift in phase which otherwise occurs in the equipment.

Either the R–C or the R–L phase-shifting circuit can be used for this purpose.

2. To obtain a shift in phase continuously variable over 360 degrees, which can be adjusted accurately to any desired degree of phase shift. Either an R–C or R–L phase-shifting circuit alone cannot be used for this purpose. However, an R–C circuit is used as a part of the Helmholtz-coil arrangement, which enables a continuously variable degree of phase shift to be obtained at a fixed voltage output.

Rotating Magnetic Field

The Helmholtz-coil phase-shifting circuit depends upon the generation of a rotating magnetic field by its primary coil assembly for its operation. Therefore, the properties of a rotating magnetic field must be considered.

The field around an electromagnet is directly proportional to the current flowing through it, and the polarity of the field is determined by the direction of current flow. When two magnetic fields are produced in the same space, the direction and magnitude of the resultant field is determined by the INTERACTION of the two fields and is the VECTOR SUM of the two fields.

The method of producing a rotating magnetic field by the action of two sets of electromagnets is shown in figure 2–42. At position (A) current flows through the vertical coils, producing a concentrated magnetic field between the upper and lower coils. The lines of force of this field point upward. At position (B) current flows through both sets of coils, producing a distributed magnetic field. The fields produced by each set of coils add to produce lines of force pointing along a diagonal upward and to the right. In other words, the field has rotated 45 degrees in a clockwise direction. At position (C) current in the vertical coils is zero, but current in the horizontal coils is at a maximum. This result is also shown by the curves of current (or field) in which the dotted line refers to the vertical coils and the solid line to the horizontal coils. The magnetic field now points horizontally to the

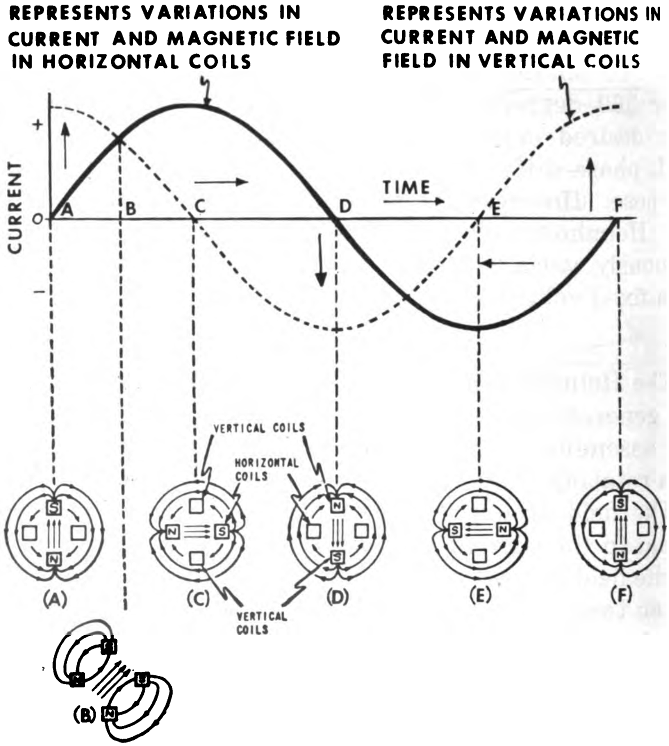

REPRESENTS VARIATIONS IN CURRENT AND MAGNETIC FIELD IN HORIZONTAL COILS
REPRESENTS VARIATIONS IN CURRENT AND MAGNETIC FIELD IN VERTICAL COILS
CURRENT
+
0
-
TIME
A
B
C
D
E
F
VERTICAL COILS
HORIZONTAL COILS
VERTICAL COILS
S
N
(A)
(B)
(C)
(D)
(E)
(F)

Figure 2–42.—Rotating magnetic field in Helmholtz coil.

right representing 90 degrees of rotation. Position (D) shows a reversal of the currents from position (A) and therefore represents a 180-degree shift in direction of the field. Likewise position (E) is a reversal of position (C), representing 270-degree rotation. Position (F) is the same as (A), showing the completion of 360 degrees rotation or one revolution of the magnetic-field flux. Thus, if alternating current is made to flow in the vertical and horizontal coils, the field can be caused to rotate uniformly.

Figure 2–43 (A) represents the condition in which the current in the vertical coils is much greater than the current in the horizontal coils. Line *A* corresponds to the field produced by the vertical coils and line *B* to the field produced by the horizontal coils. Line *C* gives the direc-

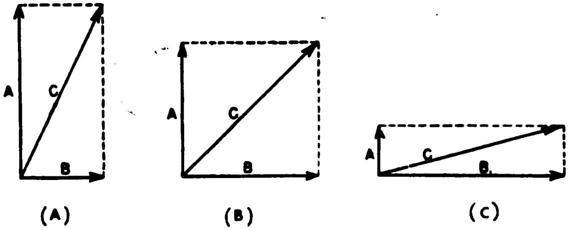


Figure 2–43.—Determination of resultant magnetic fields.

tion and magnitude of the resultant field. In figure 2–43 (B), corresponding to figure 2–42 (B), the current producing line *A* is smaller, but the current producing line *B* is larger, keeping line *C* the same length but changing its direction. In figure 2–43 (C) the current in the vertical coils is very small, but the current in the horizontal coils is so large that the resultant field, line *C*, is still the same.

Thus, if the circuit is so adjusted that the currents in the horizontal and vertical coils are always 90 degrees out of phase and their maximum amplitudes are equal, a sine-wave input produces a rotating field in the space between the coils. The direction of this field rotates through 360 degrees for each cycle of an input sine wave, while the magnitude of the field remains constant.

Helmholtz-Coil Phase-Shifting Circuit

The Helmholtz-coil phase-shifting circuit (fig. 2–44) uses two sets of coils for the primary circuit and obtains the necessary 90 degree difference in phase by means of a capacitor and resistor connected as a FIXED phase-shifting circuit in series with one set of coils. The horizontal-and-vertical-coil circuits must be carefully matched so that the maximum currents in each are equal. Since the horizontal coils are connected across the input line, the current through them lags the input voltage by almost 90 degrees. The same condition would exist in the vertical coils if they, too, were connected directly across the input terminals. But if a capacitor and an adjusting resistor are added in series with the vertical coils, this circuit can

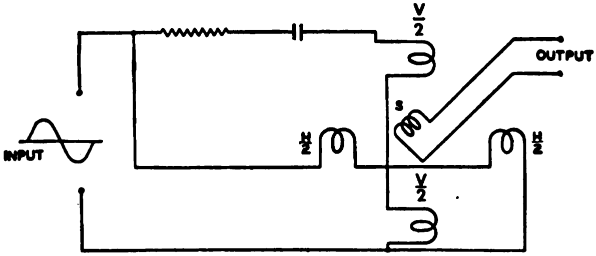


Figure 2–44.—Helmholtz-coil circuit.

be adjusted so that the current through the vertical coils is exactly 90 degrees out of phase with the current in the horizontal coils. This is the condition required to obtain a rotating magnetic field when a sine-wave voltage is applied. A secondary coil *S* is pivoted in the center of the space between the horizontal and vertical coils so that it can be rotated manually and adjusted to any desired position relative to the primary coils. The rotating magnetic field formed by the currents flowing in the primary coils induces a voltage in the secondary coil. The phase relation between the output voltage and the input voltage depends upon the physical position in which the secondary coil is placed.

In the rotating magnetic field (fig. 2–45), if some fixed point *O* is used as a reference, the magnetic lines of force point toward *O* once during each cycle. The same condition is true for any other fixed point such as *P*. However, since points *O* and *P* are separated by some angle (135 degrees in this case), the magnetic field points toward them at different times during each cycle. If the magnetic fields seen at *O* and *P* at the same instant are plotted, the resultant curves are 135 degrees out of phase (fig. 2–46). If the secondary coil is so placed that its axis passes through point *O* and the line voltage is maximum when the lines of force point to *O*, the electromotive force produced in the coil and the line voltage are maximum at the same time; in other words, they are IN PHASE. However, if the coil is rotated to the position where its axis points to *P*, a maximum electromotive force is not induced until

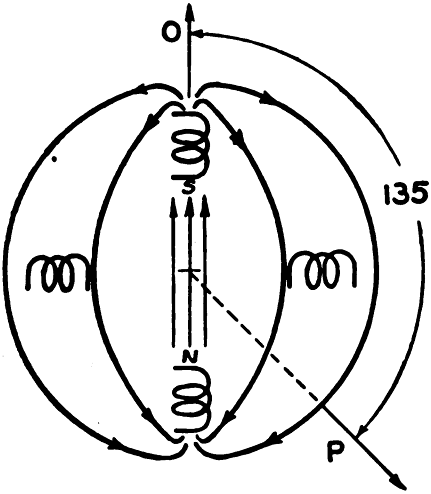


Figure 2–45.—Phase shift between line voltage and induced electromotive force.

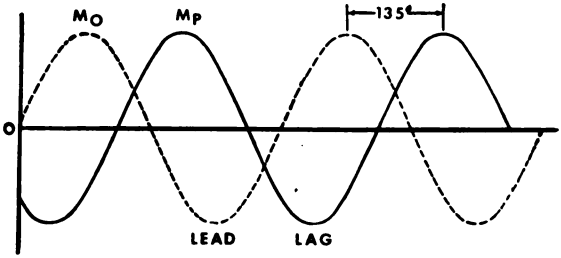


Figure 2–46.—Magnetic fields seen at O and *P* of figure 2–45.

the magnetic field rotates through 135 degrees. The electromotive force in the secondary or pickup coil and the line voltage are then 135 degrees OUT OF PHASE. Hence, any position may be selected for the axis of the pickup coil to obtain any phase relationship desired between the line voltage and the induced electromotive force. A mechanical rotation of one degree produces a phase difference of one degree between line voltage and induced electromotive force.

In order to keep the current in the secondary coil as low as practicable, a high resistance must be connected across the output of the coil. This is necessary since any appreciable current flowing in the secondary coil would produce a magnetic field which would react with the primary field and change the phase relationship between the currents in the two sets of stationary coils. The output of the circuit must be fed to a voltage amplifier.

The Helmholtz-coil phase-shifting circuit provides a means of obtaining any desired degree of phase shift, through the entire cycle of 360 degrees, by the mechanical placement of the secondary coil in the previously calibrated position.

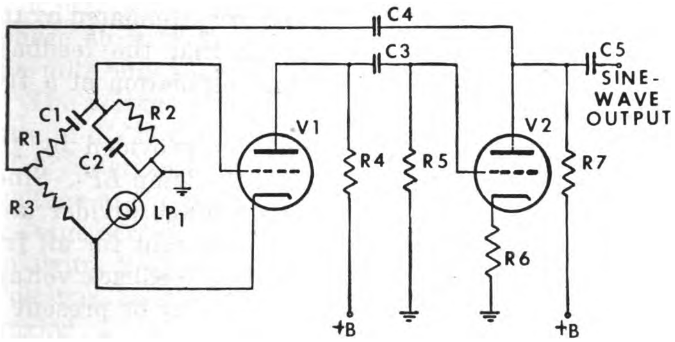
RESISTANCE-CAPACITANCE OSCILLATORS

In an R–C oscillator, the frequency is determined by a resistance-capacitance network that provides regenerative coupling between the output and input of a feedback amplifier. No use is made of a tank circuit consisting of inductance and capacitance to control the frequency.

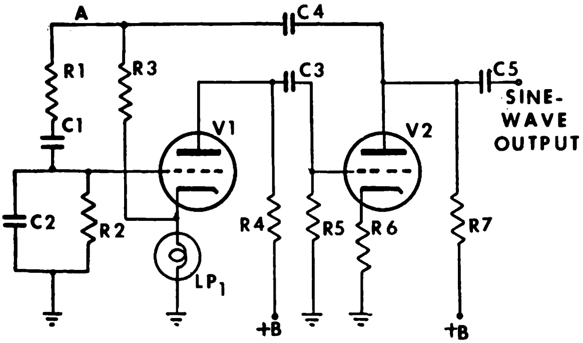
Wien-Bridge Oscillator

TYPICAL CIRCUIT.—An oscillator in which a frequency-selective Wien-bridge circuit is used as the resistance-capacitance feedback network is called a Wien-bridge oscillator. One widely used circuit for this type of oscillator is shown in figure 2–47 (A) and (B). In (A) the feedback circuit is drawn to show that the phase-shifting element of the circuit is a frequency selective bridge. However, it is simpler to use the circuit as shown in (B) for purposes of discussion since the feedback paths are shown more clearly.

ANALYSIS OF OPERATION.—Tube V_1 is the oscillator tube. Tube V_2 acts as an amplifier and inverter. Thus, even without the bridge circuit, this system oscillates since any signal that appears at the grid of V_1 is amplified and inverted by both V_1 and V_2. The voltage fed back to the grid of V_1 then must reinforce the initial signal, which causes oscillations to be set up and maintained. However, the system amplifies voltages of a very wide range of frequencies. Voltages of any frequency or of any combination of frequencies can cause oscillation. The



(A) CIRCUIT DRAWN TO SHOW BRIDGE CIRCUIT



(B) CIRCUIT DRAWN TO FACILITATE DISCUSSION

Figure 2—47.—Wien-bridge oscillator.

bridge circuit is then used to eliminate feedback voltages of all frequencies except the single frequency desired in the output.

The bridge allows a voltage of only one frequency to be effective in the circuit because of the degeneration and phase shift provided by this circuit. Oscillation can take place only at the frequency f_0 which permits the voltage across R_2, the input signal to V_1, to be in phase with the output voltage of V_2, and for which the positive feedback voltage exceeds the negative feedback voltage. Voltages of any other frequency cause a phase shift between the output of V_2 and the input to V_1 and are attenuated by the high degeneration of the circuit so that the feedback voltage is not adequate to maintain oscillation at a frequency other than f_O.

A degenerative feedback voltage is provided by the voltage divider consisting of R_3 and the lamp LP_1. Since there is no phase shift across this voltage divider, and since the resistances are practically constant for all frequencies, the amplitude of the negative feedback voltage is constant for all the frequencies that may be present in the output of V_2. A curve of the negative feedback voltage is plotted as shown in (A) of figure 2–48.

The positive feedback voltage is provided by the voltage divider consisting of R_1, C_1, R_2, and C_2. If the frequency is very high, the reactance of the capacitors is almost zero. In this case resistor R_2 is shunted by a very low reactance, making the voltage between the grid of V_1 and ground almost zero. On the other hand, if the frequency is reduced toward zero the current that can flow through either C_2 or R_2 is reduced to almost zero by the very high reactance of C_1. Therefore, the voltage between the grid of V_1 and ground falls almost to zero. At some intermediate frequency the positive feedback voltage is maximum, as shown by curve *B* in figure 2–48.

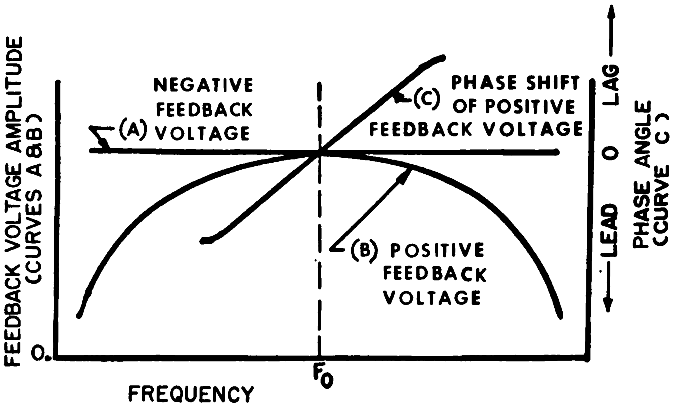


Figure 2–48.—Frequency feedback voltages in Wien-bridge oscillator.

The curve is rather flat in the vicinity of f_o, but the phase shift that occurs in the positive feedback circuit permits only a single frequency to be generated.

The voltage across R_2 is in phase with the output voltage of V_2 if $R_1C_1=R_2C_2$. If the frequency of the output of V_2 increases, the voltage across R_2 tends to lag the voltage at the plate of V_2. If the frequency decreases, the voltage across R_2 leads the output voltage of V_2. Curve C in figure 2–48 shows the phase angle between these two voltages as the frequency of the feedback voltage is varied.

The frequency at which the circuit oscillates is

$$f_o = \frac{1}{2\pi\ \sqrt{R_1C_1R_2C_2}} = \frac{1}{2\pi\ R_1C_1}$$

At this frequency the positive feedback voltage on the grid of V_1 just equals or barely exceeds the negative feedback voltage on the cathode, and the positive feedback voltage is of the proper phase to sustain oscillation. At any other frequency, the negative feedback voltage is larger than the positive, so that the resultant degeneration of the amplifier suppresses these frequencies.

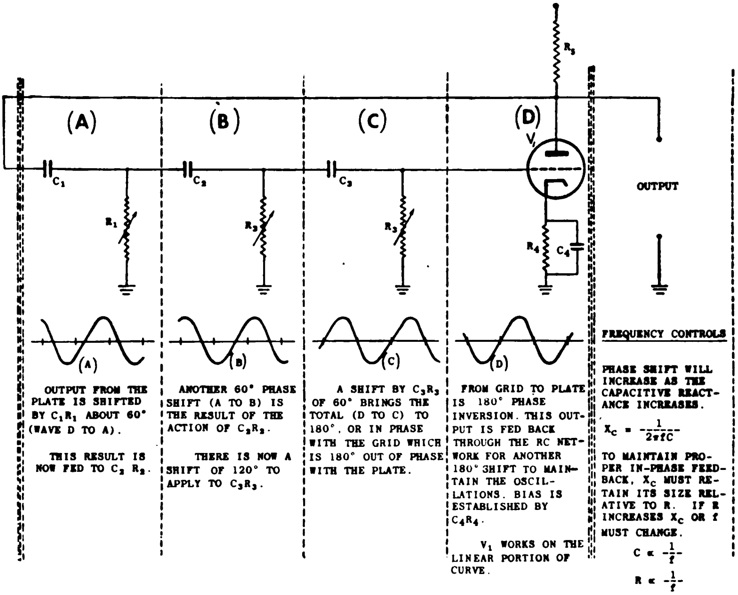
The lamp LP_1 (fig. 2–47) is used as the cathode resistor of V_1 in order to stabilize the amplitude of oscillation. If for some reason the amplitude of oscillation tends to increase, the current through the lamp tends to increase. When the current increases, the filament of the lamp becomes hotter making its resistance larger. A greater negative feedback voltage is developed across the increased resistance of the hotter lamp filament. Thus more degeneration is provided, which reduces the gain of V_1 and thereby holds the output voltage at a nearly constant amplitude. The waveform is sinusoidal only when the output from V_1 is low in amplitude. By keeping the amplitude low, the lamp, or THERMISTOR as it is sometimes called, serves also to prevent distortion of the sinusoidal waveform of the output.

APPLICATIONS OF WIEN-BRIDGE OSCILLATORS.—The

Wien-bridge oscillator has many advantages over other types of audio oscillators. For instance, it may conveniently be made to produce a wide range of frequencies. The waveshape is very nearly a true sine wave. Frequency stability is excellent. Finally, the output amplitude is nearly constant over a very wide frequency range.

Phase-Shift Oscillator

TYPICAL CIRCUIT.—The phase-shift oscillator is a special type of resistance-capacitance tuned oscillator that operates with a single tube. The circuit consists of only one amplifier tube plus a phase-shifting feedback circuit which creates a phase shift in proportion to the frequency passed through it (fig. 2–49). The standard feedback-oscillator circuit requires that the signal from the plate be shifted 180 degrees in order that reinforcing action can take place on the grid to make up for circuit losses.



OUTPUT FROM THE PLATE IS SHIFTED BY C_1R_1 ABOUT 60° (WAVE D TO A).

THIS RESULT IS NOW FED TO C_2 R_2.

ANOTHER 60° PHASE SHIFT (A TO B) IS THE RESULT OF THE ACTION OF C_2R_2.

THERE IS NOW A SHIFT OF 120° TO APPLY TO C_3R_3.

A SHIFT BY C_3R_3 OF 60° BRINGS THE TOTAL (D TO C) TO 180°, OR IN PHASE WITH THE GRID WHICH IS 180° OUT OF PHASE WITH THE PLATE.

FROM GRID TO PLATE IS 180° PHASE INVERSION. THIS OUTPUT IS FED BACK THROUGH THE RC NETWORK FOR ANOTHER 180° SHIFT TO MAINTAIN THE OSCILLATIONS. BIAS IS ESTABLISHED BY C_4R_4.

V_1 WORKS ON THE LINEAR PORTION OF CURVE.

FREQUENCY CONTROLS

PHASE SHIFT WILL INCREASE AS THE CAPACITIVE REACTANCE INCREASES.

$$X_C = \frac{1}{2\pi fC}$$

TO MAINTAIN PROPER IN-PHASE FEEDBACK, X_C MUST RETAIN ITS SIZE RELATIVE TO R. IF R INCREASES X_C OR f MUST CHANGE.

$$C \propto \frac{1}{f}$$

$$R \propto \frac{1}{f}$$

Figure 2–49.—Phase-shift oscillator.

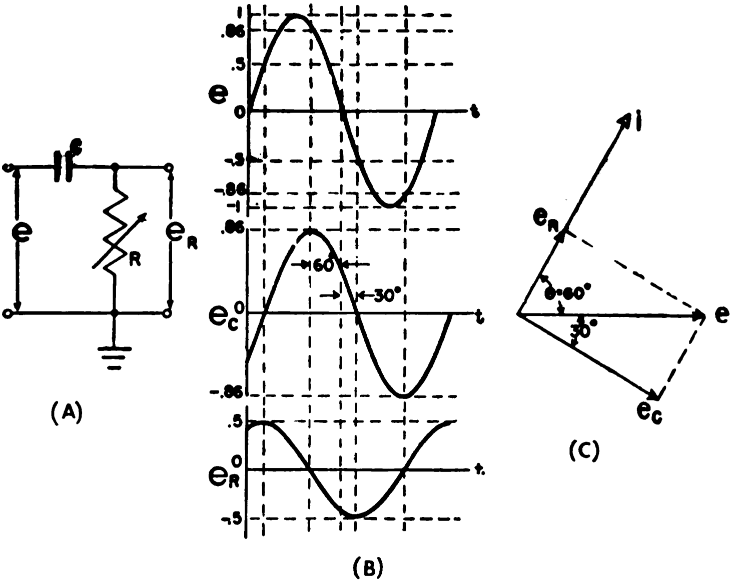


Figure 2–50.—L-section R–C phase shifter.

In the phase-shift oscillator this is accomplished by three resistance-capacitance sections.

ANALYSIS OF OPERATION.—An L-section resistance-capacitance phase shifter is shown in figure 2–50. An alternating voltage applied to this circuit causes a current to flow in the circuit. The magnitude of this current is determined by the total impedance in the circuit. Since the impedance is capacitative, the current *i* leads the impressed voltage *e* by an angle θ shown equal to 60 degrees in figure 2–50 (B) and (C). The voltage drop e_R which occurs across resistor *R* is in phase with the current that flows through it. Therefore e_R must lead the impressed voltage by an angle θ. If the output of this L-section is impressed on a second similar phase shifter, the phase of the output voltage is shifted an angle θ again. The output of this second phase shifter is then leading the first input voltage by an angle 2θ.

If the resistance of *R* is varied, the phase angle of the current flowing in the circuit also is varied. If *R* is reduced to zero, the phase angle of the current is 90 degrees, but this action is useless since there is no resistor across which to develop a useful voltage. It is not possible, then, to get a 90-degree phase shift in a single L-section of this type.

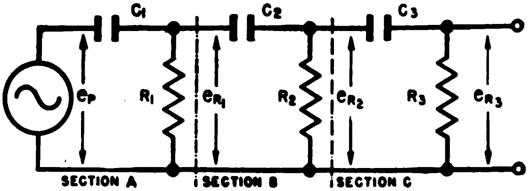
In order to obtain a phase shift of 180 degrees, three L-sections must be put in series. Because the reactance of a capacitor varies with frequency, the combination of three L-sections gives a 180-degree phase shift at only one frequency. This enables the circuit of figure 2–49 to operate at only one frequency.

The oscillations are started by any circuit change such as a plate-supply ripple or random tube noise. When a disturbance occurs, the slight change is amplified, inverted 180 degrees at the plate, and inverted another 180 degrees by the R–C network to be returned in phase to the grid of the tube for reamplification. The cumulative buildup is repeated until the tube cannot amplify further because of plate-current saturation.

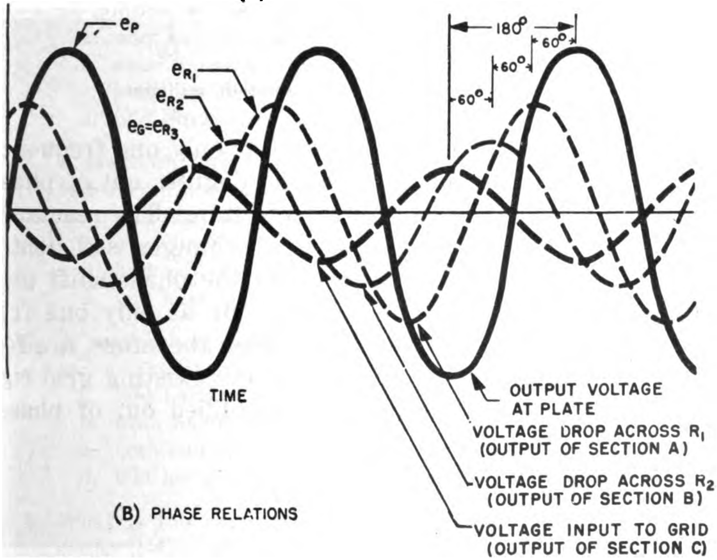
If each of the three L-sections A, B, and C (fig. 2–51) produces a 60-degree phase shift, the input voltage to the grid of the tube is shifted 180 degrees relative to the output voltage at the plate. The output of section A leads the plate voltage by 60 degrees. If the output of section A is impressed on section B, the output of this circuit leads the plate voltage by two times 60 degrees or 120 degrees. In the same way, section C adds another 60-degree shift in phase and the input to the grid therefore leads the output from the plate by 180 degrees.

The waveform of the output of the phase-shift oscillator is very nearly sinusoidal if the bias on the tube is adjusted to a value which barely allows oscillations to be maintained. The frequency stability of the circuit is then also very good.

APPLICATIONS OF PHASE-SHIFT OSCILLATORS.—The phase-shift oscillator is useful primarily in applications where a fixed frequency is desired, but the frequency can be changed by changing any of the phase-shifting capaci-



(A) SIMPLIFIED PHASE SHIFTER



(B) PHASE RELATIONS

Figure 2–51.—Phase-shifting circuit.

tors or resistors. In order to INCREASE THE FREQUENCY, either the resistance or the capacitance must be DECREASED. To DECREASE THE FREQUENCY, the resistance or capacitance must be INCREASED. The angle, that each L-section shifts the phase, is dependent on the ratio of

$$\frac{X_c}{R}=\frac{1}{2\pi\ fCR}$$

If the value of this fraction is to remain constant, either *C* or *R* must be increased when the frequency is decreased.

Another type of phase-shift oscillator is shown in figure 2–52. The bridge circuit, consisting of R_1, R_2, R_3, C_1,

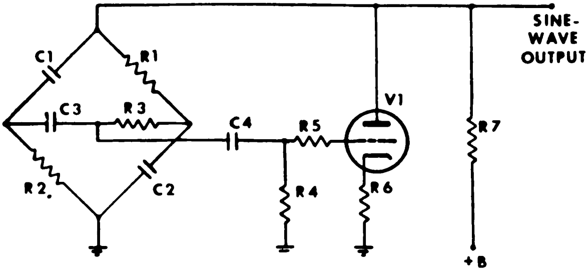


Figure 2–52.—Bridge-type phase-shift oscillator.

C_2, and C_3, is so proportioned that at only one frequency the output voltage across R_4 is 180 degrees out of phase with the voltage at the plate of the tube. The reactance of the several capacitors involved changes sufficiently with a change of frequency so that the phase shift produced is 180 degrees. This will occur at only one frequency. Voltages of other frequencies, therefore, are fed back to the grid out of phase with the existing grid signal, and are cancelled by being amplified out of phase.

QUIZ

1. A square wave is composed of a fundamental sine wave and an infinite number of
   a. odd harmonics in phase
   b. even harmonics in phase
   c. both odd and even harmonics in phase
   d. odd harmonics out of phase

2. A sawtooth waveform is composed of a fundamental sine wave and an infinite number of
   a. odd harmonics in phase
   b. even harmonics in phase
   c. both odd and even harmonics in phase
   d. odd harmonics out of phase

3. A peaked waveform is composed of a fundamental sine wave and an infinite number of
   a. odd harmonics in phase
   b. even harmonics in phase
   c. both odd and even harmonics in phase
   d. odd harmonics out of phase

4. What is the rate of charge or discharge of a capacitor in a series R–C circuit?
   a. Exponential
   b. Linear from 0% to 100%
   c. Charges/discharges 100% in one R–C time
   d. Same as in a parallel R–C circuit

5. What is the primary use of a long R–C circuit?
   a. For coupling a d-c voltage
   b. For coupling an a-c voltage
   c. For shaping
   d. To serve as a voltage divider at low frequencies

6. What is the primary use of a short R–C circuit?
   a. For coupling a d-c voltage
   b. For blocking an a-c voltage
   c. For phase shifting
   d. For waveform shaping

7. Can the universal time constant chart be used with any size resistor and capacitor?

8. An R–C circuit is considered to belong to a complex waveform when the R–C time equals
   a. $\frac{1}{10}$ the time of one period
   b. the time of one period
   c. 3 times the time of one period
   d. 10 times the time of one period

9. What is considered a short R–C circuit to a complex waveform?
   a. $\frac{1}{10}$ the time of one period
   b. 1 times the time of one period
   c. 5 times one period
   d. 10 times one period

10. What is a differentiator or integrator circuit?
   a. A coupling circuit
   b. A distortion circuit
   c. A phase splitting circuit
   d. A parallel R–C circuit

11. What is the difference in waveshape across a resistor in an R–C circuit and across an inductor in an L–R circuit?

12. Can the same universal time constant chart be used for R–C and L–R circuits?

13. How can the L–R time of a circuit be increased?
   a. Increase *L*
   b. Increase *R*
   c. Decrease frequency
   d. Increase frequency

14. What is the instantaneous value of current in a series R–C circuit after one R–C time?
   a. Maximum
   b. Minimum
   c. 63% of maximum
   d. 37% of maximum

15. What is the instantaneous value of current in a series L–R circuit after one L–R time?
   a. Maximum
   b. Minimum
   c. 63% of maximum
   d. 37% of maximum

16. What effect will increasing *R* in an L–R circuit have on resistor waveform?
    a. No effect
    b. More resemble input
    c. More distortion
    d. Increase frequency of input

17. What effect on resistor waveform will increasing *R* in an R–C circuit have?
    a. No effect
    b. More resemble input
    c. More distortion
    d. Increase frequency of input

18. What wave shape will you get across the resistor in a short R–C circuit, with a square wave applied?
    a. Square wave
    b. Peaked wave
    c. Sawtooth wave
    d. Sine wave

19. What wave shape would you get across the resistor in a long R–C circuit with a square wave applied?
    a. Square wave
    b. Peaked wave
    c. Sawtooth wave
    d. Sine wave

20. What wave shape would you get across *L* in a short L–R circuit with a square wave applied?
    a. Square wave
    b. Peaked wave
    c. Sawtooth wave
    d. Sine wave

21. What wave shape would you get across *L* in a long L–R circuit with a square wave applied?
    a. Square wave
    b. Peaked wave
    c. Sawtooth wave
    d. Sine wave

CHAPTER 3

SPECIAL CIRCUITS

LIMITERS AND CLIPPERS

The term LIMITING refers to the removal by electronic means of one extremity or the other of an input wave. Circuits which perform this function are referred to as LIMITERS or CLIPPERS.

Limiters are useful in waveshaping circuits where it is desirable to square off the extremities of the applied signal. A sine wave may be applied to a limiter circuit to obtain a rectangular wave. A peaked wave may be applied to a limiter to eliminate either the positive or the negative peaks from the output. In frequency modulation receivers, where it is necessary to limit the amplitude of the signal applied to the detection system to a constant value, limiter circuits are employed. Limiters find application as protective devices in circuits in which the input voltage to a stage must be prevented from swinging too far in the positive or the negative direction.

Series- and Parallel-Diode Limiters

SERIES-DIODE LIMITING.—The characteristics of a diode are such that the tube conducts only when the plate is at a positive potential WITH RESPECT TO THE CATHODE or, in other words, when the cathode is negative with respect to the plate. If the cathode is held at ground potential, the plate need only be positive with respect to ground for current to pass through the diode. A positive potential may be placed on the cathode, in which case the tube does not conduct until the voltage on the plate rises above an equally positive value. Likewise, the cathode may be held at a negative potential, and the tube conducts while the plate is positive and continues to conduct while the plate is at a negative potential which is less negative than the cathode. As the plate becomes more positive with respect to the cathode, the current through the tube increases and the plate-to-cathode resistance decreases rapidly from an open circuit to an average value on the order of 500 ohms.

The series-diode limiter (fig. 3–1) is commonly used to limit the positive half of a sine wave. The rectifying characteristics of the diode are utilized so that it may be considered as a switch. This is justified if the value of *R* is very large as compared to the resistance of the diode when conducting. Thus, in figure 3–1 the output voltage remains at zero throughout the positive half cycle of the input since the diode switch is open and no current flows through *R*. During the negative half cycle, on the other hand, the cathode is negative with respect to the plate and the tube conducts. The switch is closed and the output voltage developed across *R* follows the applied voltage and, neglecting the very small drop across the tube e_p, is essentially equal to it.

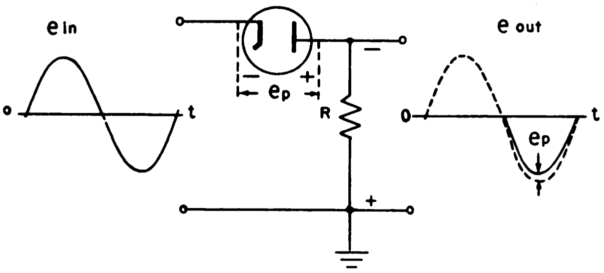


Figure 3–1.—Series diode used to limit positive signals.

In a similar manner the same circuit, with the diode connections reversed, may be used to limit the negative swing of the input voltage. This application is illustrated in figure 3–2. The diode switch is closed during the positive swing of the input voltage and is open during

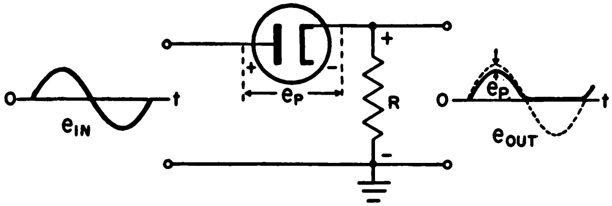


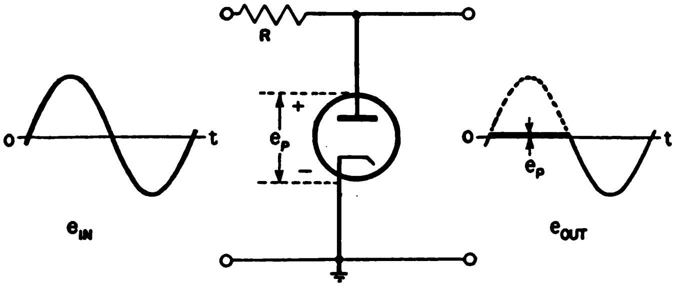
Figure 3–2.—Series diode used to limit negative signals.

the negative swing. Thus a voltage is developed across *R* during the positive half cycle only.

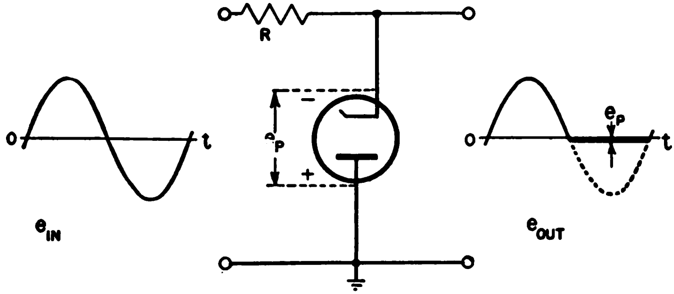
PARALLEL-DIODE LIMITING.—An alternate method of employing diodes in limiter circuits is shown in (A) and (B) of figure 3–3. The tubes in the two circuits are connected in parallel with the load, which is assumed to be a very high impedance so that the output current is negligible.

In figure 3–3 (A) the diode is connected so as to limit the positive signals at approximately ground potential. Since the cathode is held at ground potential, the diode conducts throughout the entire positive half cycle. Current flows through the tube and through the series resistor *R*. As *R* is large as compared to the plate-to-cathode resistance of the diode, essentially the entire input voltage is developed across *R* and the output voltage is only the very low voltage drop across the diode e_p. This may be a negligible positive voltage depending on the ratio of *R* to the diode resistance. On the negative swing of the input the diode does not conduct; thus no current flows through *R* and the output voltage equals the input.

In figure 3–3 (B) the plate of the diode is held at ground potential so that the tube does not conduct during the positive half cycle. Thus the output voltage equals the input. During the negative half cycle of the input voltage the cathode is negative with respect to the plate and the diode conducts. The diode current flows through the series resistance *R* across which essentially the entire input voltage is developed. The output voltage is



(A) POSITIVE LIMITING



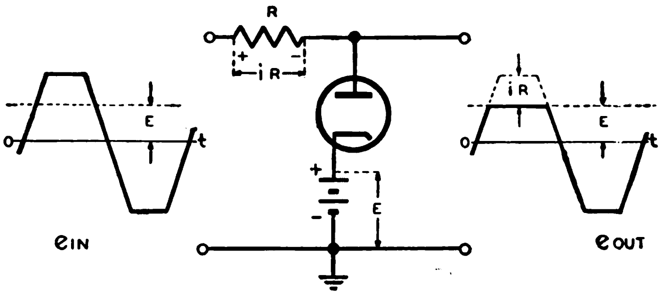
(B) NEGATIVE LIMITING

Figure 3–3.—Parallel-diode limiter circuits.

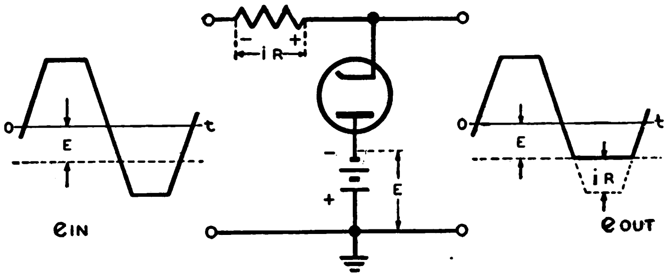
limited to the very low voltage drop across the tube. This low negative voltage as a rule may be neglected, and the outputs may be considered as being limited at essentially ground potential as a result of the switching action of the diode.

BIASING THE LIMITER.—An input voltage can be limited to any desirable positive or negative value by holding the proper diode electrode at that voltage by means of a battery or biasing resistor. Two such circuits are shown in figure 3–4.

The cathode of the diode in figure 3–4 (A) is more positive than the plate by the value of *E* when no signal is applied at the input. As long as the input voltage remains less positive than the battery voltage *E*, the diode acts as an open switch and the output equals the input.



(A) POSITIVE LIMITING

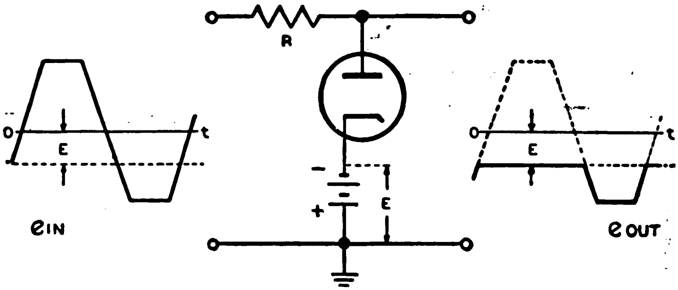


(B) NEGATIVE LIMITING

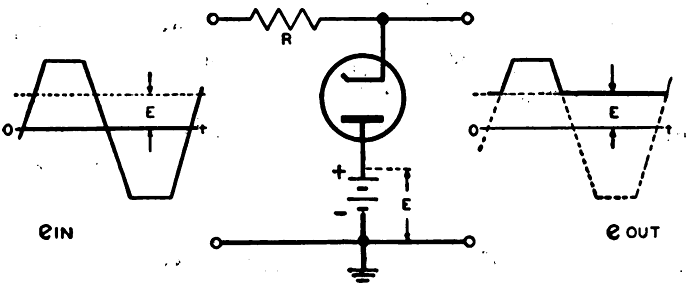
Figure 3–4.—Parallel diodes limiting above and below ground.

If the input increases to a value greater than *E*, the diode conducts and behaves as a closed switch which effectively connects the upper output terminal to the positive terminal of the battery. Thus during this portion of the input cycle, the output voltage equals *E* and the difference between e_{in} and *E* appears as an *IR* drop across the resistor *R*, neglecting e_p.

The plate of the diode (fig. 3–4 (B)) is negative by the value of battery voltage *E*. Thus, as long as the input is positive or is less negative than *E*, the diode is an open switch and the output voltage e_{out} is equal to the input. When the input becomes more negative than *E*, the diode conducts and effectively connects the upper output terminal to the negative terminal of the battery. During this portion of the input cycle, e_{out} equals *E* and



(A) NEGATIVE PEAKS RETAINED



(B) POSITIVE PEAKS RETAINED

Figure 3–5.—Parallel-diode limiters which pass peaks only.

the difference between e_{in} and E appears as an IR drop across R.

It is sometimes desirable to pass only the positive or negative extremity of a waveform on to a succeeding stage. To accomplish this the parallel-diode limiters shown in figure 3–5 can be employed. In (A) the entire portion of the input waveform above the negative potential E causes the diode to conduct, thus producing an output voltage which varies between the negative level of E and the negative extremity of the input. In (B) the diode conducts during the entire portion of the input waveform which is below the positive potential of E. The output voltage then varies between the positive level of E and the positive extremity of the input waveform.

In either case the difference between the value of *E* and e_{in} during the time the diode conducts is represented by the *IR* drop across the series resistor *R*.

CRYSTAL-DIODE LIMITING.—Although most circuits use diode vacuum tubes in limiter circuits, the crystal diode or silicon crystal rectifier is becoming increasingly popular for limiting. The crystal diode usually consists of a small piece of silicon and a pointed tungsten wire contactor. It operates by virtue of the fact that it passes little current from the crystal to the wire because of high resistance, but permits current to flow readily from the wire to the crystal when an a-c voltage is applied to its input terminals. Since, like a diode vacuum tube, it conducts more current in one direction than in the other, the crystal assembly often is called a CRYSTAL DIODE. Not only is the crystal almost as efficient as the vacuum tube, but it requires no heater voltage and is entirely free of interference from the power supply frequency. It may be used in any of the diode limiter circuits previously described. The one illustrated in figure 3–6 is for use in limiting or clipping the positive cycle of the input waveform.

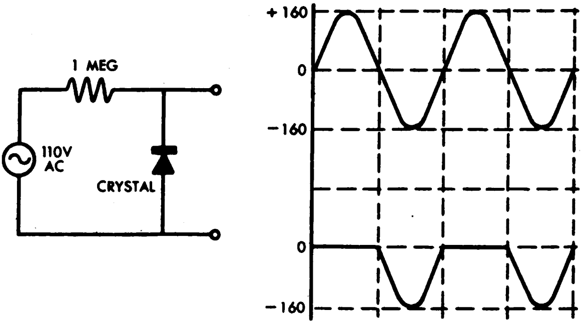


Figure 3–6.—Positive crystal limiter.

DOUBLE-DIODE LIMITING.—It is possible to limit both amplitude extremities of an input waveform at any desirable level by placing two diodes in parallel in the limiter circuit. In figure 3–7, the diode V_1 is made to

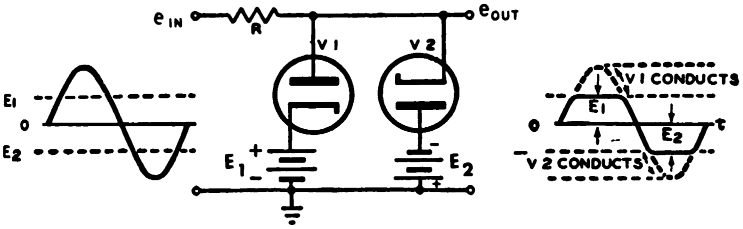


Figure 3–7.—Double-diode limiter circuit.

conduct whenever the input voltage e_{in} reaches a higher positive value than E_1, thus limiting the positive half cycle to the value of E_1. The diode V_2 is made to conduct whenever the input reaches a higher negative value than E_2, thus limiting the negative half cycle to the value of E_2. The circuit represents a simple method of producing a satisfactory square-wave output with a sine-wave input voltage.

Amplifier Tubes as Limiters

GRID LIMITING ACTION.—The grid-cathode circuit of a triode, tetrode, or pentode may be employed as a limiter circuit in exactly the same way as the plate-cathode circuit of the diode limiter illustrated in figures 3–3 and 3–4. By inserting a series grid resistor, as shown in figure 3–8, which is very large compared to the grid-to-cathode resistance when grid current flows, essentially the entire positive half cycle of the input voltage is limited to the voltage level of the cathode. For example, the grid-to-cathode resistance may drop from an infinite value, when the grid is negative with respect to the cathode, to a value on the order of 1,000 ohms, when the grid attempts to become positive with respect to the cathode. If a 1-megohm resistor is placed in series with the grid, the drop across the 1,000-ohm R_{GK} is negligible as compared to that which is developed across the 1-megohm resistor by the flow of grid current.

The grid-limiter circuit shown in figure 3–8 (A) is held normally at zero bias. During the positive portion of the input signal the grid attempts to swing positive.

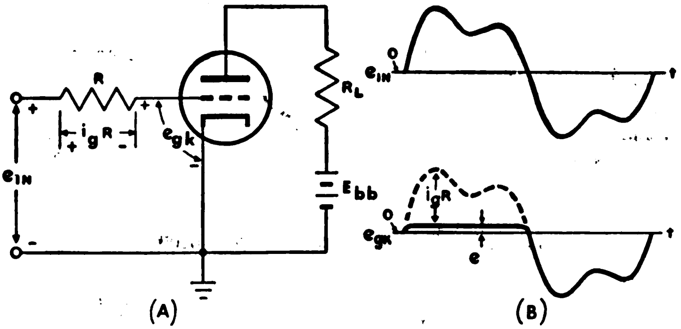


Figure 3–8.—Unbiased grid-limiter circuit.

Grid current flows through the resistor *R*, developing an i_gR drop of such polarity as to oppose the positive input voltage. Since the full input voltage must appear as the sum of the drop across *R* and R_{GK}, the larger *R* is with respect to R_{GK}, the nearer the voltage on the grid is limited to that of the cathode. The i_gR drop may be considered as an automatic bias developed during the part of the input which causes grid current to flow.

Alternate circuits for limiting the positive peaks of the input voltage are shown in figures 3–9 and 3–10. In figure 3–9 (A) the tube is biased by the negative potential *E* supplied to the grid, with the cathode returned to ground. No grid current flows until the input signal e_{in} rises sufficiently to equal and effectively remove the biasing voltage *E*. Any further rise of e_{in} drives the grid positive with respect to the cathode, and grid current flow limits the signal on the grid by virtue of the i_gR drop across *R*.

In figure 3–10 (A) bias is developed between grid and cathode by the flow of plate current through the cathode resistance R_K which is bypassed by the large capacitor C_K. The grid is held normally at ground potential and thus is negative with respect to the cathode. Any positive signal e_{in} must drive the grid positive by an amount equal to the value of *E* before the biasing effect of R_K is removed. This is shown by e_g waveform in figure 3–10. A further rise of the input voltage produces grid current

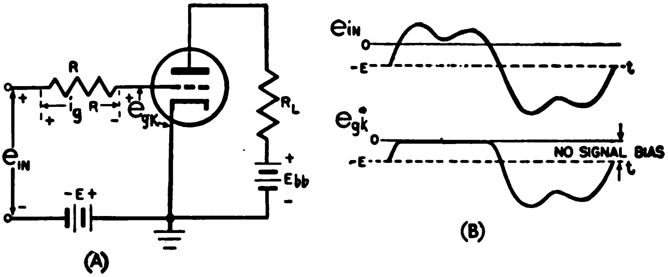


Figure 3–9.—Grid limiter with cathode grounded and grid biased by negative voltage.

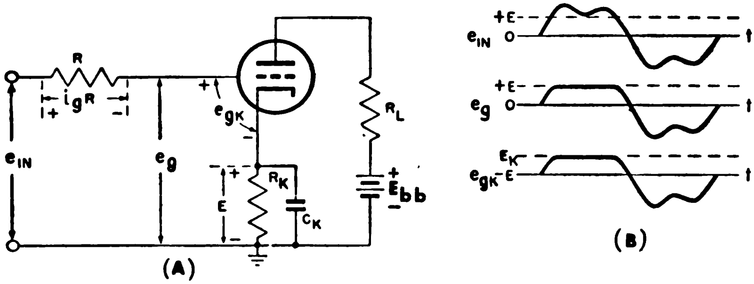


Figure 3–10.—Grid limiter with automatic cathode bias.

which results in the limiting of the voltage at the grid. The grid-to-cathode voltage e_gK in figure 3–10 (B) illustrates the effect of R in preventing the grid from swinging appreciably above the cathode potential.

SATURATION LIMITING.—Whenever a series-limiting resistor is used in the grid circuit, the grid cannot be driven to an appreciable positive voltage and, despite the positive amplitude of the input voltage, the maximum plate current which flows is that determined by the plate supply E_{bb} and the resistance of the plate circuit at zero bias. Thus the minimum plate voltage is determined by the limiting action in the grid circuit. These plate-current and plate-voltage relationships are shown in figure 3–11.

The grid-limiting resistor may be omitted, if the input signal comes from a low-impedance high-power source,

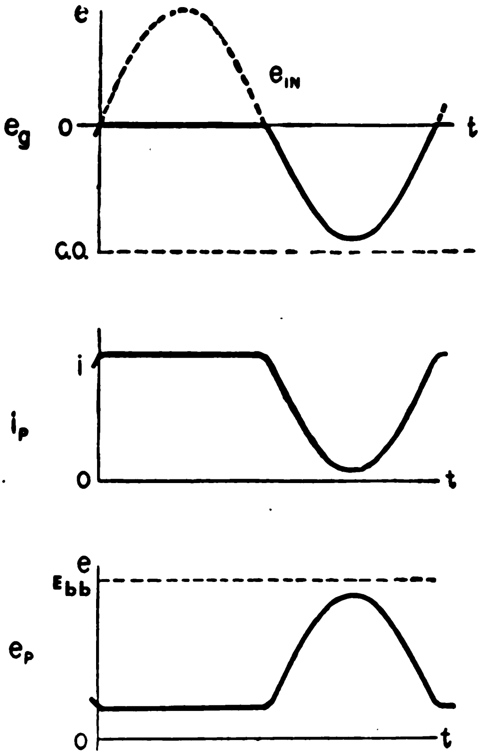


Figure 3–11.—Effect of grid limiting on plate current and plate voltage.

and limiting in the plate circuit may still be realized. This is due to plate-current saturation and is usually referred to as SATURATION limiting. Plate-current saturation should not be confused with emission saturation since in tubes using oxide-coated cathodes there is no definite saturation value of emission current.

By using a large value of plate-load resistance R_L and a low plate-supply voltage E_{bb}, saturation limiting may be produced by a relatively low amplitude of positive grid voltage. In any case, however, the plate current can never exceed the value E_{bb}/R_L. In an actual circuit some small positive voltage must remain on this plate to attract the electrons from the cathode, and the saturation plate current never quite equals E_{bb}/R_L. In other words there remains across the tube a low voltage drop when the plate current is at saturation, since the plate-to-cathode resistance at saturation does not decrease to zero.

In figure 3–12 (A) the i_p vs. e_p characteristic of a triode is used to illustrate the effect of saturation limiting on the plate voltage. In (B) the input signal applied to the grid, which is normally at zero bias, is not of sufficient amplitude to drive the tube to cutoff on the negative swing but causes the plate current to saturate on the positive swing. The dotted extension of the loadline describes the tube during the positive position of the input cycle. The maximum plate current shown in (C) cannot exceed the value E_{bb}/R_L no matter how high the amplitude of the positive grid signal, and is actually slightly

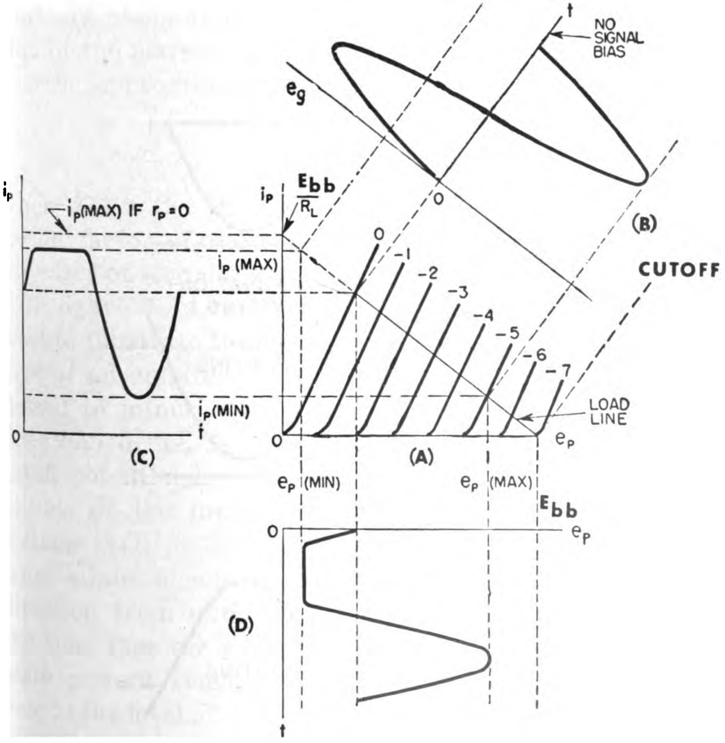


Figure 3–12.—Tube characteristics (i_p vs. e_p) illustrating saturation limiting.

less because of the low-saturation plate resistance which remains in series with the load. In (D) the maximum plate current defines the lowest value to which the plate voltage can fall. During the remaining portion of the input cycle, the grid controls the flow of plate current which in turn determines the shape of plate-voltage waveform.

The results of saturation limiting are similar to those of grid limiting in that the negative-going portion of the plate voltage is affected. These are compared in figure 3–13. Saturation limiting has the advantage of producing an output wave of greater amplitude, but it has the disadvantage of requiring considerably more power to drive the grid.

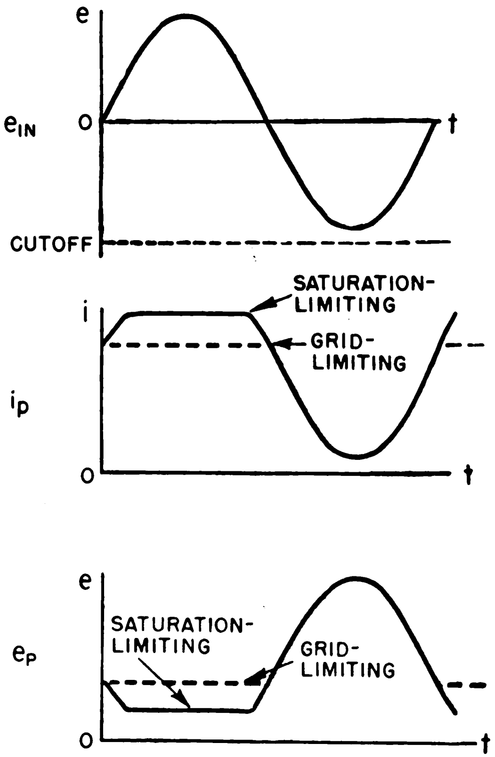


Figure 3–13.—Comparison of grid and saturation limiting.

CUTOFF LIMITING.—Electron current through a vacuum tube can flow only from cathode to plate and not from plate to cathode. Therefore, plate current cannot become a negative value. When the grid is driven to cutoff, the plate current is decreased to zero and remains at zero during the time the grid is below cutoff. Since no current flows through the plate circuit when the tube is cut off, there is no voltage developed across the load resistance, and the plate is maintained at the full value of the plate-supply voltage. Thus, a type of limiting is achieved in which the positive extreme of the plate waveform is flattened as a result of driving the grid beyond cutoff.

The cutoff voltage may be defined as the negative voltage with respect to the cathode, to which the grid must be driven in order to prevent the flow of plate current. For any given type of tube this voltage level is a function of the plate-supply voltage and in the case of TRIODES may be approximated by the expression

$$E_{co} = \frac{E_{bb}}{\mu}$$

where E_{bb} is the plate-supply voltage and μ is the amplification factor of the tube. This relation is NOT valid in the cases of tetrodes and pentodes.

In figure 3–14 the i_p vs. e_p characteristic of a triode is used to illustrate the limiting effect caused by driving the grid of an amplifier beyond cutoff. The grid normally is biased to minus 5 volts by the steady drop across R_K. (See (A) of fig. 3–14.) The value of E_{bb} is such that the cutoff potential E_{co} is minus 7 volts. The maximum amplitude of the input voltage is 4 volts; thus the grid voltage ((C) of fig. 3–14) swings in a positive direction from minus 5 volts to minus 1 volt and in a negative direction from minus 5 volts to minus 9 volts. During the time that the grid voltage remains below cutoff, the plate current remains at zero and the plate voltage is held at the level of E_{bb}.

A combination of grid limiting and cutoff limiting may

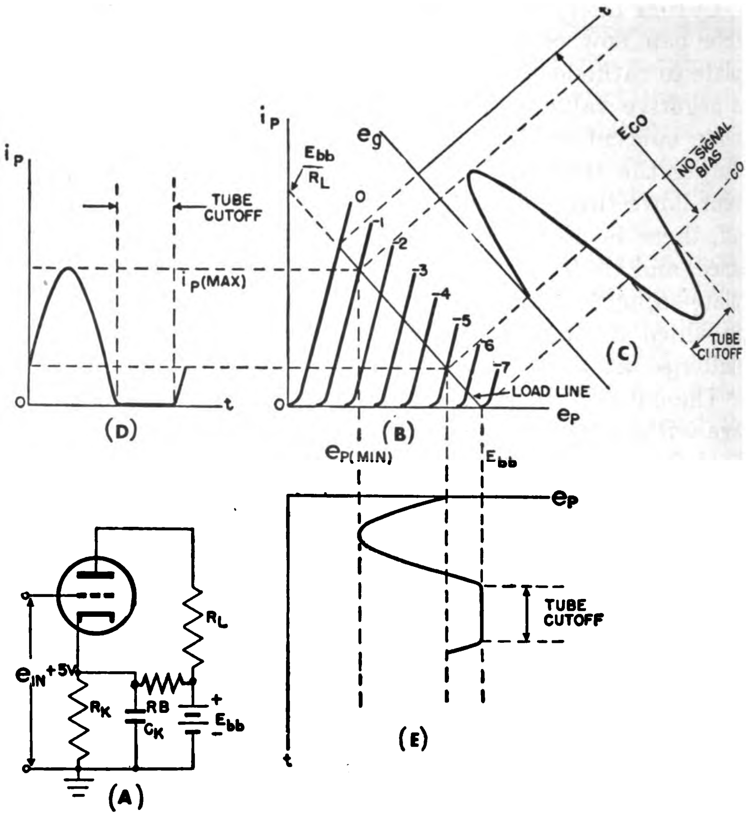


Figure 3–14.—Squaring top of plate voltage by cutoff limiting.

be employed in an amplifier circuit to produce a square wave from a sine wave. (See fig. 3–15.) The amplitude of the input voltage is sufficiently high to hold the grid beyond cutoff for the greater part of the negative swing. The grid resistor *R* is on the order of 1-megohm and limits the grid voltage essentially to zero during the positive swing.

THE OVERDRIVEN AMPLIFIER.—An amplifier circuit in which SATURATION LIMITING is employed in conjunction with CUTOFF LIMITING to produce a rectangular wave from a sine wave is generally known as an OVERDRIVEN

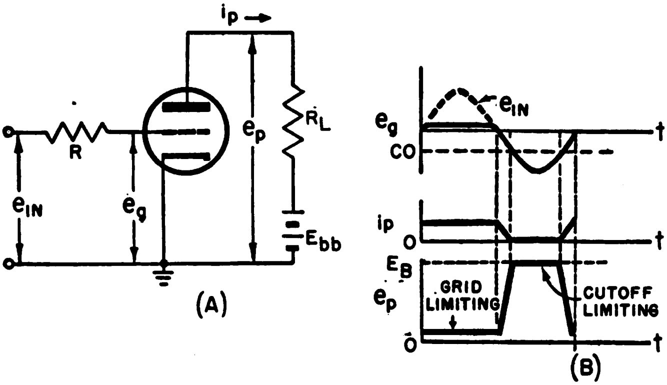


Figure 3–15.—Formation of a square wave by grid and cutoff limiting.

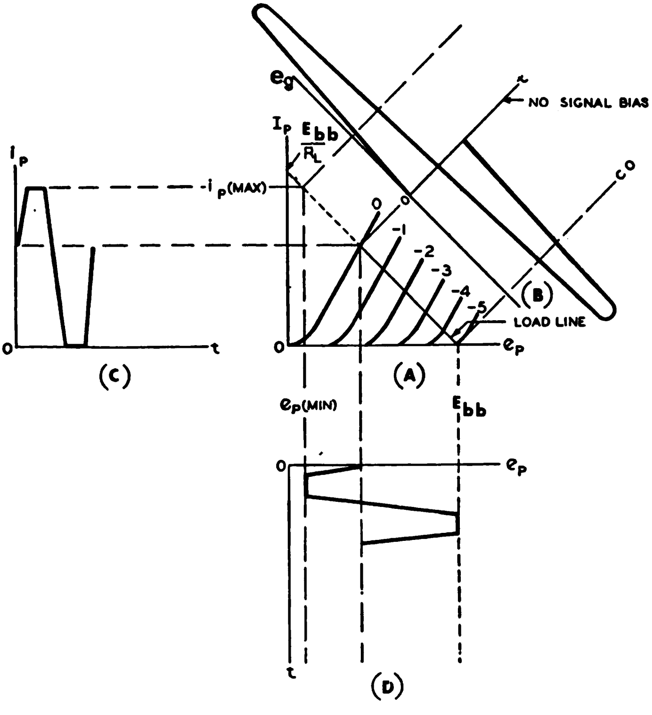


Figure 3–16.—Formation of a square wave by saturation and cutoff limiting.

AMPLIFIER. The driving circuit for such an amplifier should have a relatively low-output impedance and be capable of delivering power since considerable current is drawn during the positive swing of the grid. The value of the load resistor R_L is made as large as practicable for the plate-voltage supply available.

Figure 3–16 illustrates the grid-voltage, plate-current, and plate-voltage relationships as determined by the i_p vs. e_p charactertistics and the circuit constants of the overdriven amplifier.

Typical Application of Limiters

THE SQUARER AND PEAKER CIRCUIT.—The squarer and peaker circuit uses the principles discussed in the preceding limiter circuits. Each part of this circuit is used in several electronic circuits, and the entire circuit is used in some radar sets.

Notice the simple squarer and peaker circuit in (A) of figure 3–17. The first tube is an overdriven amplifier which squares the input sine waveshape and amplifies it according to principles of triode limiting. The 0.002-μf. capacitor and 500-K resistor form a short time-constant combination which peaks the square wave. The grid of the second triode is biased so that only a positive or negative amplified peak appears in the output. Because of this limiting action it is called a limiter or clipper stage.

CIRCUIT ANALYSIS.—In analyzing the peaker circuit, first inspect the circuit diagram. The first stage is a triode which looks like an amplifier stage except that it is not biased. However, there is a 1-megohm grid-limiting resistor in its grid circuit. This stage applies its output through capacitive coupling to the second triode, a stage which uses a 100-K load resistor and fixed bias of 50 volts. Bias in this stage is provided by a 250-K and 50-K voltage divider across the positive 300-volt supply.

Now consider the operation of the peaker. As the input voltage goes from zero to a positive voltage, the grid of the first triode is made positive with respect to its cathode. Immediately, grid current flows through the cathode-to-grid resistance of the tube and the 1-megohm resistor. The internal resistance of the cathode-to-grid path is less than a thousand ohms; therefore, the voltage developed by grid current divides at a ratio of 1,000 to 1 across the tube and the plate-load resistor. At the positive peak of the input voltage peak (163 volts), the grid voltage is about a tenth of a volt. The grid waveshape shows this voltage as a flat line at the zero level. (The dotted line is the applied voltage; the solid line is the actual grid voltage.) The plate voltage of the first triode during this time is rather low because the plate current is unlimited by the grid. To determine its exact value, construct a loadline on the characteristic curves as follows: First, calculate the maximum current by dividing the plate supply of 300 volts by the load resistance of 50K. Thus, the maximum current is 300/50K, or 6 ma. Draw a line across the curves from the 300-volt zero current point on the bottom to the 6 ma. zero voltage point along the current scale. This line is the loadline.

The first use of the loadline is to find the plate voltage for zero grid voltage. At the intersection of the zero grid-voltage curve and the loadline, drop down to the plate-voltage scale at the bottom. The point shows that the plate voltage is 55 volts. During the positive swing of applied voltage, the grid actually goes slightly positive. This is shown by the amplified curve of the plate-voltage change which is shown below the grid-voltage curve. At the start of the half cycle, the curve is at 55 volts, but drops below that slightly as the grid goes slightly positive, then returns to 55 volts at the end of the half cycle.

Next, consider the input voltage during the negative half cycle. First, however, remember that the grid does not draw current when it is negative. Therefore, when no current flows through the 1-megohm resistor, there will be no voltage drop across it. Thus, the full input voltage will appear at the grid during the entire negative

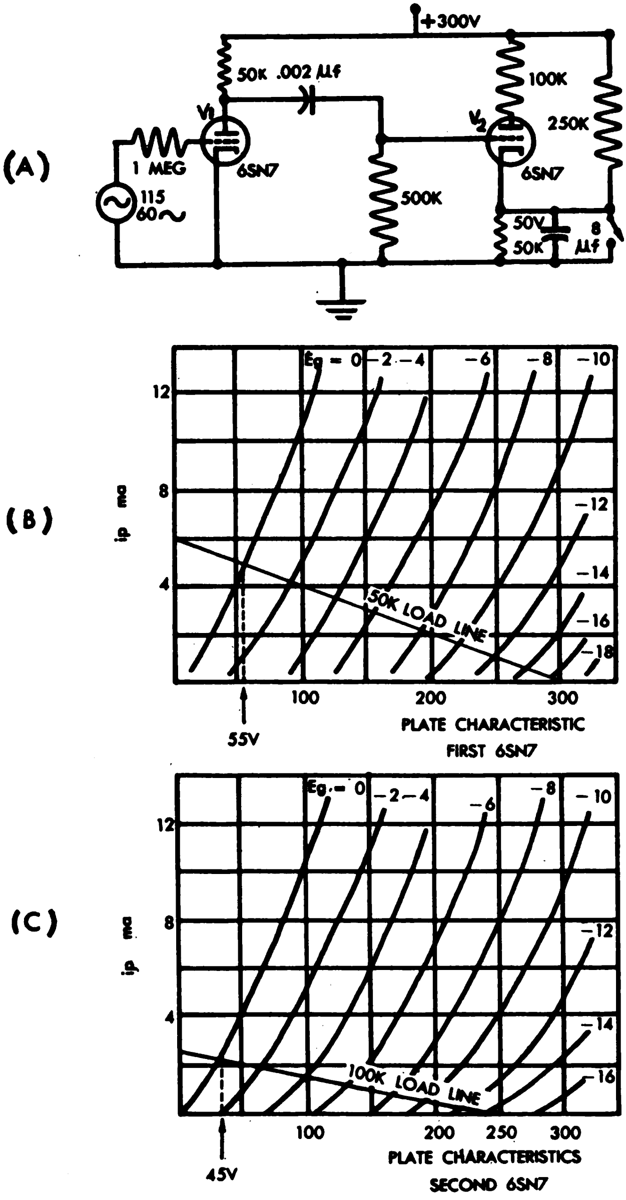
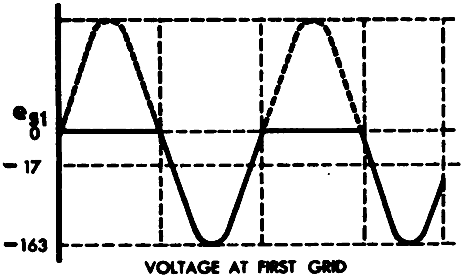
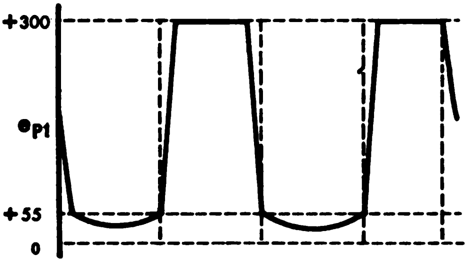


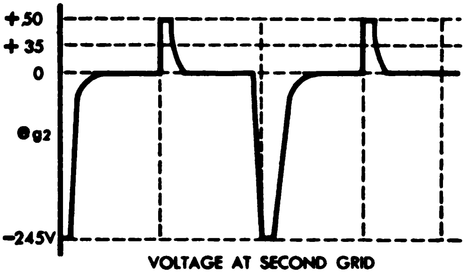
Figure 3–17.—Squarer and peaker circuit.



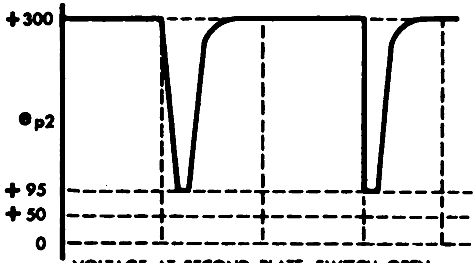
VOLTAGE AT FIRST GRID



VOLTAGE AT FIRST PLATE



VOLTAGE AT SECOND GRID
SWITCH OPEN



VOLTAGE AT SECOND PLATE SWITCH OPEN

Figure 3–17.—Squarer and peaker circuit. Continued.

half cycle as it is shown in the grid-voltage curve. On glancing at the loadline, notice that it crosses the zero current axis at the end of the minus 17-volt grid curve. Thus, when the grid voltage reaches minus 17 volts, the plate current will have decreased to zero. The grid curves show that this value is reached very quickly during the negative half cycle. The grid voltage continues to swing with the applied voltage through minus 163 and back to zero. All this time the grid voltage is more than 17 volts negative, and the plate current remains at zero. As the grid voltage starts to become negative, the decrease of current through the 50-K load resistor causes the plate voltage to increase toward the 300-volt supply voltage. The change is very rapid because the grid voltage crosses the cutoff value in less than 1/100 of the cycle. Therefore, the plate voltage rises to plus 300 volts during the entire time the grid voltage is below the cutoff value on the curve. As it comes above cutoff, the plate voltage again drops to 55 volts. And again, the grid-voltage change from minus 17 to zero is very steep, and accordingly the plate-voltage change is exceedingly fast. The long period of time that the grid is below cutoff and the rapid changes cause a square wave to be developed at the plate of the first triode.

The square-wave output of the first stage is applied to an R–C circuit consisting of the 0.002-μf. capacitor and 500-K resistor in the grid of the second tube. The voltage across the resistor becomes the grid voltage. At the start of the first half cycle the square wave is negative. Since the capacitor is initially uncharged, the application of this negative-going voltage to it immediately causes current to flow through the R–C circuit. With no drop across the uncharged capacitor, the full voltage appears across the resistor. The third curve labeled e_{g2} shows the entire voltage change occurring at the second grid.

Suppose for a moment you examine the conditions in the second tube when this grid voltage is applied to it. The cathode is 50 volts positive with respect to ground.

The actual plate voltage is 300 minus the 50 volts on the cathode, or 250 volts. Notice the loadline on the lower set of characteristic curves. This line is drawn from the 250-volt supply-voltage zero current point to the 250/-100K, or 2.5 ma. zero voltage point. On referring to this line, notice that the plate current becomes zero when the grid voltage is minus 15 volts. With 50 volts bias on this tube, there will be no current until more than plus 35 volts is applied to the grid. Therefore, the negative swing of 245 volts applied to the grid will not change the plate current from its zero value. But also note that this waveshape is not like the square wave applied because the R–C time constant is short. The total time the voltage is applied is 1/120 or 0.00833 second, while the R–C is 0.002 μf. × 0.5 megohm or 0.001 second, a ratio of about 8 to 1. In other words the capacitor is small and becomes charged during the very first portion of the square wave. As it becomes charged, the current through the 500-K resistor becomes lower and lower until after 5 time constants or 0.005 second, it drops to zero. The resistor voltage stays in step with this decreasing current so the voltage drops exponentially to zero. The grid voltage stays at zero for the duration of the half cycle or until the reversal of the square wave occurs.

At the end of the first half cycle the voltage at the first triode changes 245 volts in the positive direction. Again this change immediately appears across the resistor. Nothing happens, however, until the voltage reaches 35 volts. Then current starts to flow in the tube. As the voltage increases from 35 volts to 50 volts, the actual grid-to-cathode voltage at the second tube rises from minus 15 volts to zero. Now the plate waveshape for this second tube can be examined. During the entire previous half cycle, the tube was at cutoff and the plate voltage remained at plus 300 volts. But now current is flowing, and the voltage drops to that value of plate voltage for zero grid voltage. This value can be determined again from the loadline. Find the intersection of the zero grid voltage line with the loadline and proceed downward to the plate-voltage scale. It reads 45 volts in this curve. This might seem inconsistent inasmuch as the plate-voltage curve in (C) of figure 3–17 indicates that the plate voltage is 95 volts. But remember that the voltage is being indicated with respect to ground. Therefore, the 45 volts across the tube must be added to the bias voltage to obtain the total voltage from plate to ground. Now return to the second grid. As the square wave gets above 50 volts, the grid becomes positive and starts to draw current. So you see that the positive-going voltage on the second grid levels off at zero. Only a small part of the remaining voltage rise of 245 volts ever appears across the grid to cathode.

This is due to the fact that the grid current flows through the 0.002-μf. capacitor and 50-K load resistor, and this combination of the 50-K resistor and the high reactance of the small capacitor acts as a grid limiter to divide the total voltage. The small grid-to-cathode resistance of about 500 ohms will have a very small voltage drop across it. The 500-K resistor is still in the circuit but so much of the current flows through the grid itself that the 500-K resistor can be practically disregarded during this time. The low resistance and small capacitor form a very short R–C circuit, so the capacitor will be charged quickly. The voltage at the grid then drops back to zero with respect to ground which is 50 volts negative with respect to the cathode. So the triode conducts only during the brief period of time that the grid is above cutoff. The plate voltage stays at plus 300 volts most of the time. A negative pulse is thus generated for each cycle of the original input sine wave. Its amplitude is 300 minus 95 or 205 volts.

When the switch in this circuit is closed, the circuit will generate a positive pulse for each cycle of the sine-wave input. As shown in figure 3–18, when the switch is closed in the actual circuit it shorts out the 50-K bias resistor and its parallel capacitor, and connects the 300-K resistor to ground. These three circuit elements become ineffective and an equivalent circuit may be drawn by

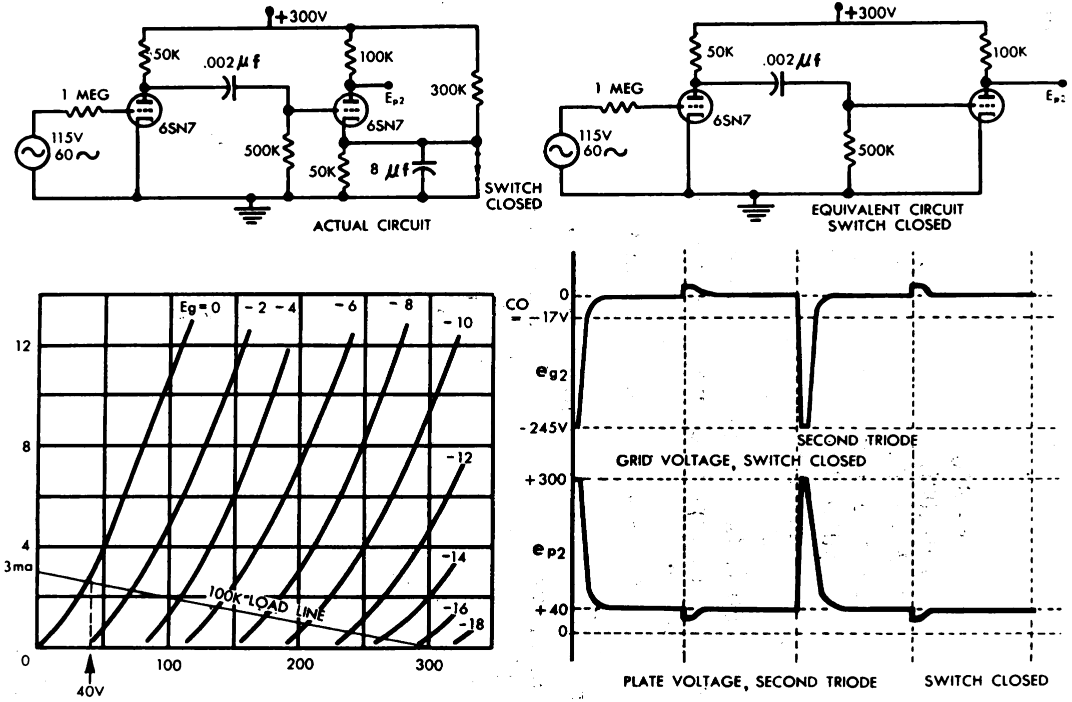


Figure 3–18.—Squarer and peaker with switch closed to form negative peak.

omitting these elements. Now there is no bias on the second triode, and the full 300-volt supply voltage appears across the tube and its load resistor. To evaluate the current and voltages, it is necessary to draw a new loadline as shown in the characteristic curves. This line extends from the 300-volt zero current point at the bottom to the 300/100K or the 3 ma. zero voltage point at the side. With no bias, the grid voltage is zero, and the plate voltage under this condition is 40 volts.

On the negative half cycles the square wave will be peaked in the same manner as before. But on the positive peaks, grid current flow occurs at zero instead of the plus 50 volts as before. Similarly, as in the previous case, the high impedance circuit of the 0.002-μf. capacitor and 50-K load resistor prevents the grid from getting more than a fraction of a volt positive during positive half cycle.

The grid remains at zero value except when the input voltage is negative-going: Therefore at zero grid voltage the plate current being high, the plate voltage which varies inversely with the plate current remains at a constant value of 40 volts. At the close of the first half cycle, the voltage is negative-going. Here the grid voltage changes rapidly from zero to the cutoff value, and the plate voltage rises sharply from 40 volts to plus 300 volts. The grid voltage becomes negative by 245 volts, and the resulting plate voltage levels off at the supply-voltage level of 300 volts. However, the rapid charge of the 0.002-μf. capacitor causes the grid voltage to rise immediately to the cutoff value and beyond. As the grid voltage continues to rise to zero, the plate voltage drops back to 40 volts. If you study the plate-voltage waveshape, you notice the fairly constant low plate voltage with positive peaks with each cycle of input voltage. The positive peak voltage at the second grid only raises the grid a fraction of a volt above zero, thus causing the plate voltage to dip slightly at each of these points. Thus, with this circuit it is possible to obtain either positive or negative peaked pulses by the flip of the switch.

DIRECT-CURRENT RESTORERS OR CLAMPERS

A circuit which holds either amplitude extreme of a waveform to a given reference level of potential is called a CLAMPING CIRCUIT. The terms D-C RESTORER and BASE-LINE STABILIZER also are commonly used. Such circuits are divided roughly into two classifications. The first, DIODE AND GRID CLAMPING, clamps EITHER amplitude extreme and allows the waveform to extend in only one direction from the reference potential. The second, SYNCHRONIZED CLAMPING, maintains the output potential at a fixed level until a synchronizing pulse is applied, at which time the output potential is allowed to follow the input. At the end of the synchronizing pulse the output voltage is returned immediately to the reference level.

To demonstrate the utility of clamping circuits, a brief review of the action of coupling networks is useful. In coupling between stages in electronic circuits a coupling capacitor almost always must be used to keep the high positive d-c plate potential of the first tube isolated from the grid of the second tube. It is desirable that only the VARYING component of the plate potential be transmitted to the grid as a signal varying above and below some fixed reference level. If the lower end of the grid-leak resistor is grounded, the signal varies above and below ground. If a biasing potential is employed, the signal applied to the grid varies above and below this d-c bias voltage.

In class A operation, the latter condition is desirable. The biasing potential is adjusted to the center of the class A range and the varying potential is kept within the limits of this range. The center of the grid swing is fixed, and the amplitude variation of the grid voltage directly affects the amplitude of the plate-voltage swing. This is exactly the condition desired in class A operation (fig. 3–19 (A)).

In other circuits, however, the waveform swing must be entirely above or entirely below the reference voltage, instead of alternating on both sides of it (fig. 3–19 (B)

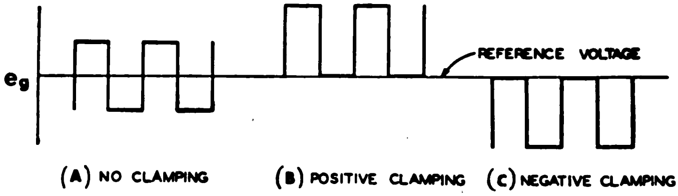


Figure 3–19.—Grid voltage variation with respect to definite reference potential.

and (C)). For these applications a clamping circuit is used to hold either the positive extreme or the negative extreme of the waveform to the desired level.

The output of an ordinary R–C coupling network is alternating in character swinging about the average voltage level of the applied waveform. After the coupling capacitor charges to the average applied voltage, any decrease in applied voltage causes the output voltage of the R–C network to swing negative. Any increase above the average causes the output voltage to swing positive. Now if the capacitor can be made to charge, to say to the MINIMUM applied voltage and no more, any swing has to be in the positive direction. The output voltage varies between zero and some positive value, depending on the amplitude of the input signal. If, on the other hand, the capacitor can be made to charge to the MAXIMUM applied voltage and remain charged to that level, any swing necessarily is in the negative direction. The output voltage varies between zero and some negative value, depending on the amplitude of the input signal. In the first case the bottom of the output waveform is clamped to zero (fig. 3–19 (B)), and in the second case the top of the output waveform is clamped to zero (fig. 3–19 (C)).

Positive and Negative Diode Clampers

The simplest type of clamping circuit utilizes a diode in conjunction with the ordinary R–C coupling circuit. Consider the case in which the capacitor voltage is maintained at the minimum applied voltage (fig 3–20). In order to understand the action of this circuit, the following significant point should be kept in mind: If the cathode of a diode is made negative with respect to the plate, or the plate positive with respect to the cathode, electrons flow from cathode to plate and the tube constitutes a low resistance. On the other hand, if the cathode is made positive with respect to the plate or the plate negative with respect to the cathode, no current flows and the tube may be considered an open circuit.

The plate-voltage variation of a circuit producing a square-wave output (fig. 3–21) is typical of the kind of input applied to the clamping circuit of figure 3–20. In this clamping circuit, capacitor *C* charges gradually through the high resistance *R*. After a period of time, depending on the R–C time constant, the charge on the capacitor reaches 50 volts, the base of the input waveform. The problem is to maintain the charge at this value in spite of the tendency of the capacitor to charge to a higher level when the applied voltage goes to plus 150 volts.

Assuming that a steady voltage equal in magnitude to that at point *A* in figure 3–21 has been applied for some time, the capacitor in figure 3–20 may be considered as being charged to 50 volts. During the time between *A* and *B* the charge on the capacitor is equal to the applied voltage and no current flows. Then at point *B* the applied potential suddenly increases to plus 150 volts. Since it is impossible for the charge on the capacitor to change instantaneously, the difference between the plus 150 volts applied and the 50 volts across the capacitor must appear across *R*. The difference of 100 volts becomes the output voltage.

The fact that a voltage appears across *R* (fig. 3–20) indicates that a current is flowing through it. This current flow adds to the charge on *C*. Ordinarily the R–C time constant would be very long and the actual charge added to *C* would be extremely small. For simplicity, assume that the 150-volt potential is applied for a time equal to 1/10 *RC* or from point *C* to point *D*. Since the

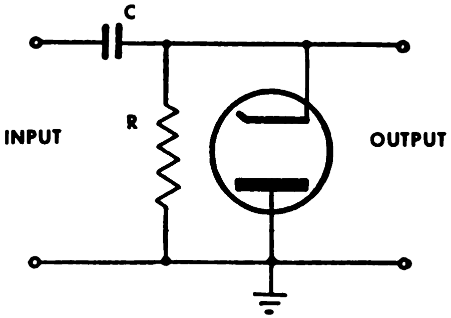


Figure 3–20.—Positive clamping circuit.

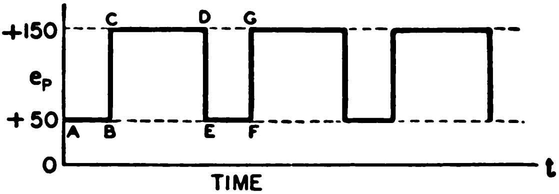


Figure 3–21.—Typical plate voltage applied to clamping circuit.

cathode of the diode is positive with respect to the anode, the tube is in effect an open circuit and *R* is a high resistance. During a time equal to 1/10 *RC* the charge on the capacitor increases exponentially by 10 percent of 100 volts, or 10 volts, making the total charge 60 volts. During the same time the drop across the resistor decreases exponentially by 10 volts to a value of 90 volts, leaving the sum of e_R and e_C still equal to the applied potential of 150 volts.

Now at point *D*, the applied voltage suddenly drops back to 50 volts, but the capacitor is charged to 60 volts. This would leave an output voltage across *R* of 10 volts negative with respect to ground—a condition it is hoped to avoid. In order for the output to return to zero very quickly, the capacitor must discharge the extra 10 volts through a very short R–C path.

In figure 3–20 the cathode of the diode is connected to the high side of *R* and the plate is grounded. Any output voltage which is negative with respect to ground

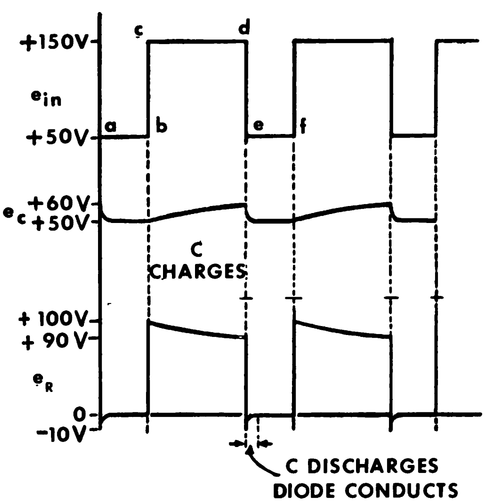


Figure 3–22.—Waveforms e_C and e_R of diode-clamping circuit with grounded plate for input signal e_{in}.

makes the cathode negative with respect to the plate. In this state the diode conducts and becomes, in effect, a very low-resistance discharge path for the capacitor until the charge is again equal to the applied voltage and the output voltage returns to zero. These conditions are illustrated in figure 3–22.

To illustrate the operation of the positive-clamping circuit further, it is assumed that the waveform shown in figure 3–23 is applied to the clamping circuit of figure 3–20. Since at point *A* the input voltage is zero, the output voltage is zero and remains so until point *B* is reached. At this time the input voltage drops suddenly to minus 100 volts at point *C*. Since the capacitor cannot change its charge instantaneously, the output voltage across *R* also drops suddenly to minus 100 volts. Since the cathode of the diode is 100 volts negative with respect to the plate, the tube conducts heavily, charging the capacitor very rapidly through a short R–C until the capacitor voltage becomes equal to the applied voltage. At this time the output voltage has returned to zero and the diode becomes nonconducting. As long as the input

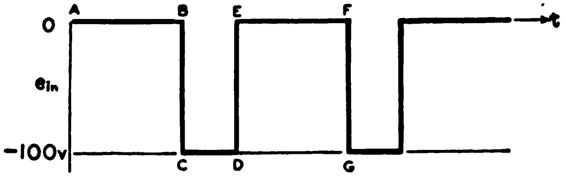


Figure 3–23.—Typical voltage applied to diode-clamping circuit.

remains at minus 100 volts, from point *C* to point *D*, the output voltage remains at zero potential.

At point *D*, the input voltage changes back to zero, a rise of 100 volts in the positive direction (minus 100 to 0). This rise produces a rise of 100 volts (0 to plus 100) across *R*, as the capacitor again cannot change its charge instantaneously. The capacitor must now discharge very slowly as the diode is nonconducting and the high-resistance path through *R* must be utilized.

Assuming again a time for discharge from points *E* to *F* of 1/10 *RC*, the voltage across the capacitor at *F*, and thus the output voltage, decreases to 90 volts, since the input is zero. At point *F* the input signal again drops to minus 100 volts. Instantaneously the output across *R* goes to minus 10 volts (input minus e_C). The diode conducts quickly, returning the charge on the capacitor to 100 volts and the output to zero. These results are shown in figure 3–24. Note that no portion of the wave-

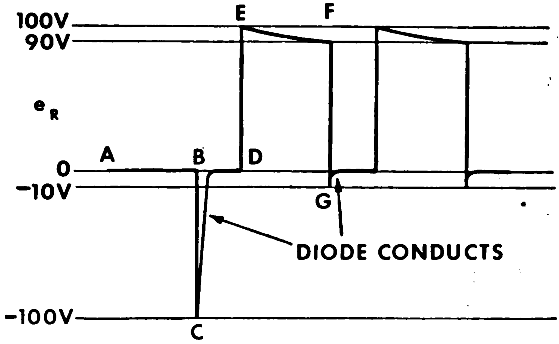


Figure 3–24.—Output waveform of diode-clamping circuit with grounded plate and input signal as shown in figure 3–23.

form is lost after the first cycle. The function of the clamping circuit is merely to shift the reference voltage from the top of the waveform to the bottom.

Figure 3–25 illustrates a diode-clamping circuit capable of causing the output waveform to vary between some negative value and the zero reference voltage. The only difference between this circuit and the one shown in figure 3–20 is in the manner in which the diode is connected. In this case the plate is connected to the output terminal while the cathode is grounded. Therefore in this condition the tube conducts whenever the plate rises above ground.

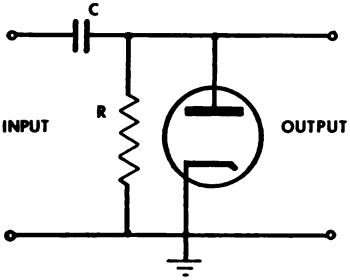


Figure 3–25.—Negative-clamping circuit.

The input signal (fig. 3–21) is now applied to this negative-clamping circuit. When the input is at point *A* the plate of the diode becomes positive and the capacitor quickly charges up to plus 50 volts through the short R–C path. When this occurs the output voltage drops to zero. Then at point *B* the input voltage suddenly rises 100 volts (from plus 50 volts to plus 150 volts). Since the capacitor charge cannot change instantaneously, the 100-volt change appears across *R*, sending the plate of the diode to plus 100 volts with respect to the cathode. The capacitor again is presented with a short R–C charging path and charges very rapidly, bringing the output voltage back to zero quickly. At point *D* the input signal suddenly drops 100 volts, but capacitor *C* is charged to 150 volts and cannot change instantaneously. The 100-volt drop therefore appears across *R*, sending the output from zero to minus 100 volts. The input voltage drops from plus 150 volts to plus 50 volts and the output drops from 0 volts to minus 100 volts. During the time from *E* to *F* the capacitor can be expected to discharge a small amount through the high-resistance path of *R*. At point *G* this slight loss of charge is replaced quickly as the plate of the diode goes positive momentarily. The output signal is shown in figure 3–26.

In the practical circuit the size of the resistance *R* is great enough to make negligible the amount of distortion of the output caused by the slight charging of the capacitor.

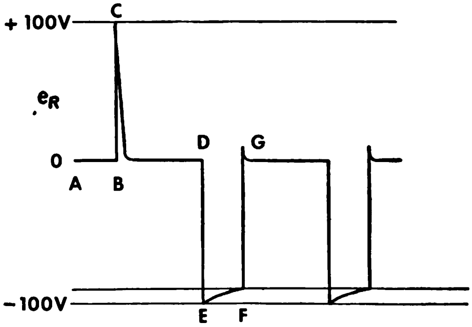


Figure 3–26.—Output waveform of diode-clamping circuit with grounded cathode and input signal as shown in figure 3–21.

GRID CLAMPING.—The function of clamping may be performed at the grid of an ordinary triode or pentode as well as in a diode. Any element of a vacuum tube, if made positive with respect to the cathode, attracts electrons from it. On the other hand, any element made negative with respect to the cathode repels electrons and has no current flow. Thus the grid of a tube, connected as shown in figure 3–27, acts as the plate of a diode and produces the same clamping action as the circuit of figure 3–25. Any tendency for the grid to go positive causes grid current to flow, charging capacitor *C* to the applied potential.

CLAMPING ABOVE OR BELOW GROUND POTENTIAL.—Al-

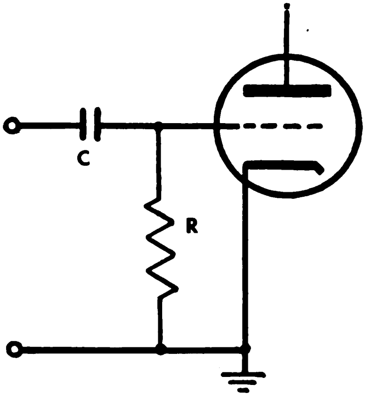


Figure 3–27.—Grid-clamping circuit.

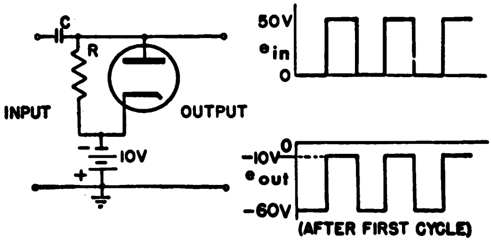


Figure 3–28.—Clamping circuit which establishes reference potential of minus 10 volts.

though circuits previously discussed clamped one extreme of the input signal to zero potential, actual circuits need not be limited to this one reference potential. Figure 3–28 illustrates the means of clamping the upper extreme of an input signal to minus 10 volts with respect to ground. The same principle may be applied to clamp either voltage extreme to any reference potential.

Synchronized Clamping

A more versatile clamping circuit is shown in figure 3–29. This arrangement keeps the bias on V_3 constant except for the time during which the clamping tubes V_1 and V_2 are held beyond cutoff by a synchronizing pulse. During the period of conduction of V_1 and V_2,

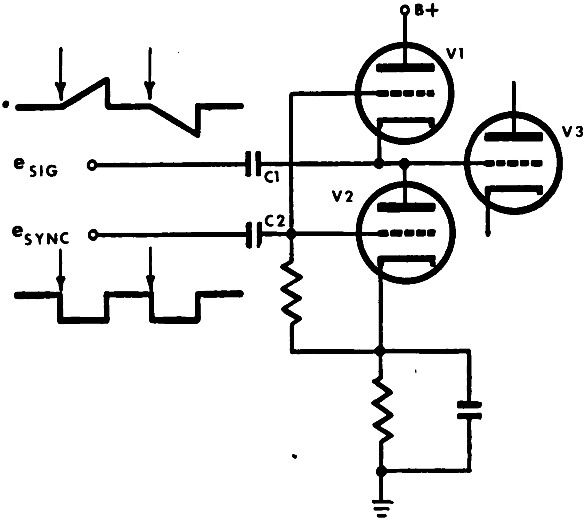


Figure 3–29.—Synchronized clamping circuit.

these tubes may be considered as part of a voltage-divider network for the purpose of placing a definite voltage on the grid of V_3. Actually the circuit is slightly more complicated than the simple voltage-divider explanation would imply.

In figure 3–29, while V_2 is operating at zero (grid-to-cathode) bias and may be considered therefore as a simple resistor, V_1 is biased by the drop across V_2. The reason is that the cathode of V_1 is connected directly to the plate of V_2 and the grid of V_1 is tied directly to the grid of V_2, which in turn is connected effectively to the cathode of V_2, so long as no synchronizing potential is applied.

Assume that for some reason the potential at the grid of V_3 tries to rise. The voltage at the grid of V_3 and the voltage across V_2 are identical. Thus a rise in the grid voltage of V_3 results in an increase in the bias on V_1, which makes it a higher resistance than before, bringing the voltage at the plate of V_2 back to normal. In like manner a decrease in the voltage at the grid of V_3 means a decrease in the bias of V_1, which decreases the resistance of V_1, bringing the voltage at the grid of V_3 back to normal. As long as V_1 and V_2 conduct, the voltage at the grid of V_3, and thus the plate current of V_3, are held constant by the voltage-divider action of V_1 and V_2. V_1 is the variable resistance which controls the fraction of the B-plus voltage that appears across V_2.

The synchonizing pulse is applied as a negative rectangular wave which drives V_1 and V_2 beyond cutoff for the desired length of time. With tubes V_1 and V_2 cut off, the grid of V_3 is left free to follow any changes in input-voltage amplitude. Capacitor C_1 has no path through which to discharge except by way of the very high resistance of the insulating material used for tube bases and sockets. Thus the voltage at the grid of V_3 follows exactly the input voltage. At the end of the synchonizing pulse V_1 and V_2 again conduct, returning the voltage at the grid of V_3 quickly back to the reference potential.

Application of Clamping Circuits

In practice, clamping usually is encountered in sweep circuits. If the sweep voltages do not always start from the same reference point, the trace itself does not begin at the same point on the screen each time the cycle is repeated, and is therefore jittery or erratic. If a clamping circuit is placed in the grid circuit of the final sweep amplifier, the voltage from which the sweep signal starts can be regulated by adjusting the d-c voltage applied to the clamping circuit. An arrangement such as that shown in figure 3–28 insures, in the case of a negative-going sweep voltage, that the waveform applied to the deflection plate varies in a negative direction from the d-c voltage level applied to the cathode of the diode.

A typical radar PPI (Plan Position Indicator) deflection amplifier that uses the synchronized clamping circuit just discussed is shown in figure 3–30. The equipment makes use of four deflection amplifiers (two vertical and two horizontal). A dual triode clamping tube is used with each deflection amplifier. Figure 3–30 shows only

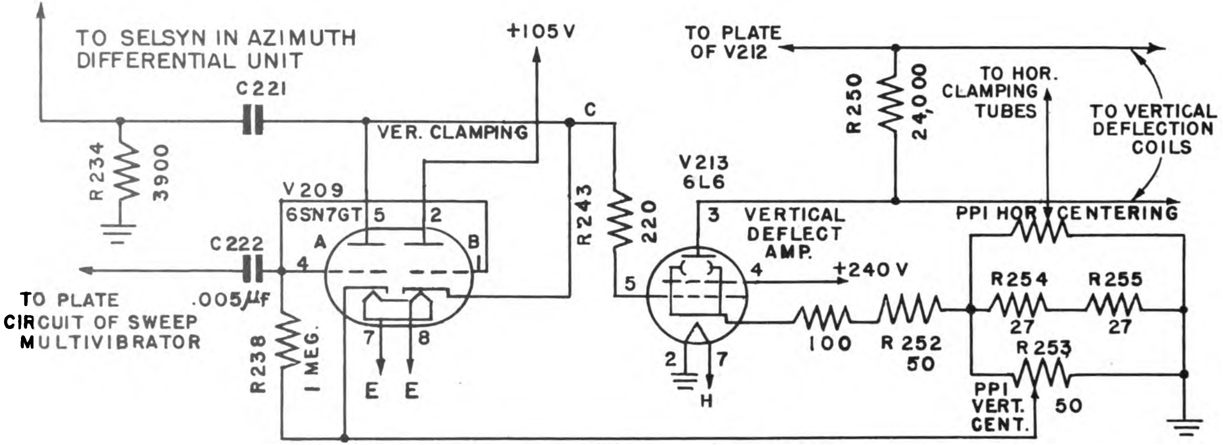


Figure 3–30.—Clamping tube circuit.

one clamping and one deflection amplifier. The other three circuits are identical with this one. Notice in figure 3–30 how the cathode resistor of the clamping tube *V*209*A* is now a variable resistor *R*253. The resistor is used to center the start of the PPI sweep vertically. A similar adjustment for horizontal centering is contained in the horizontal clamping circuits.

An application of these clamping circuits may be found in radar equipment where GROUND STABILIZATION is added to the system. Ground stabilization is used for the purpose of causing the start of the PPI sweep to be displaced. This displacement is governed by the speed and direction of the aircraft containing this radar equipment. One result of this action is that ground targets will remain in a fixed position on the scope. Secondly, all targets that are moving, such as ships or aircraft, will move across the scope in their true direction and at their true speed. This displacement of the start of the sweep is accomplished by varying the bias on the clamping tubes. The varying bias is derived from the auxiliary ground stabilization equipment.

TIME BASE GENERATORS

A time base generator circuit is a circuit that generates the voltage which causes the spot to move across the CRT screen at a constant rate. The distance along the trace formed by the moving spot is the basis of establishing time intervals. When recurrent voltage changes are superimposed on the trace, it is possible to measure both time and frequency of a radio signal, and in radar to measure the distance of targets. Commercial test oscillographs employ a thyratron tube in the time base generator circuit because of the simple way in which the frequency can be changed. In radar sets the time base generator circuit usually employs a high vacuum tube.

Review of the Thyratron Sweep Generator

The movement of the spot in the CRT may be caused either by a voltage change on the deflection plates or by a

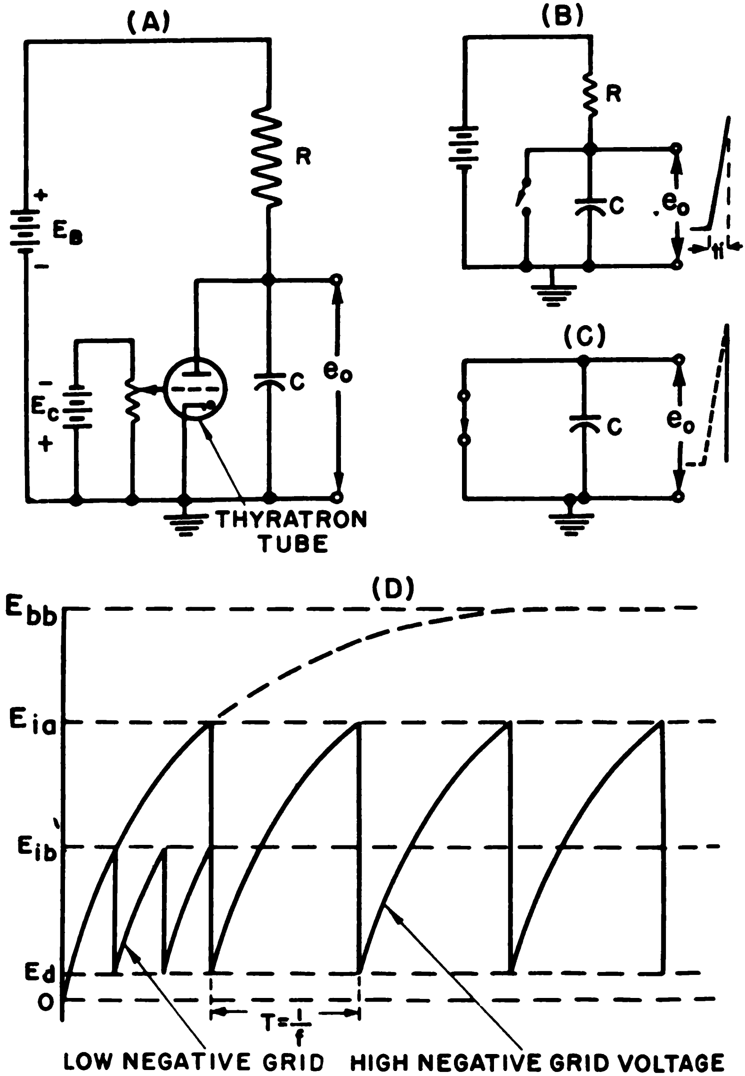


Figure 3–31.—Thyratron sweep generator.

current change in the deflection coils. The chief requirement is that the voltage (or current) must change linearly with time.

Although both the R–C and L–R circuits consume time during a voltage change, the R–C circuit is the one usually preferred as a means of generating a changing voltage. As previously mentioned, an R–C circuit contains a d-c voltage source, a series resistor, and a capacitor. When these parts are connected, the capacitor will start to charge; but since the resistor limits the current, the charge will not be instantaneous, but at a rate which is exponential. The charging curve is quite linear in the very beginning, and the capacitor charging process is usually stopped before the slope of the charging curve changes appreciably. After the charging stops, the capacitor discharges, and the events just described repeat and the capacitor recharges.

Most time base generators use this very same charging and discharging process. Any circuit differences that exist result largely from the methods employed for high-speed switching of the capacitor from charge to discharge. One device used for this switching action is the vacuum tube. The circuit in (A) of figure 3–31 shows a gas triode or thyratron, a special type of electron tube, employed for providing the necessary switching operation since this tube has certain special characteristics which make switching occur automatically. In the circuit the tube is in parallel with the capacitor. When the circuit is turned on, there is no charge on the capacitor and the tube acts like an open circuit due to the negative cutoff voltage on the control electrode, or grid. The capacitor charges along the first curve in (D) of figure 3–31. As the voltage increases, a potential is reached where the voltage across the tube is too high for the control electrode to keep the tube cut off and a small plate current starts to flow. This plate current ionizes the gas in the tube, and before the voltage can increase any further a sudden high plate current flows. The plate voltage for the tube is provided by the charged capacitor.

Forcing a current through the tube causes discharge of the capacitor. Below a certain potential, the tube does not remain ionized, and when the capacitor is discharged down to this potential, the gas in the tube de-ionizes, and the grid regains control of plate current.

At this point it becomes an open circuit again and permits the capacitor to recharge. For further study refer to the equivalent circuits; the charge of the capacitor in (B) of figure 3–31 and the circuit during discharge in (C).

The curve for several charges and discharges in (D) of figure 3–31 shows that the voltage increases to the ionizing potential E_{ia} or E_{ib} and then drops very abruptly to E_d which is the de-ionizing potential. The slow rise is due to the size of R and C. The rapid drop in voltage forms an R–C curve, but the resistance of the ionized tube is so small that the capacitor is discharged almost immediately, causing the curve to be practically vertical. The voltage at which ionization occurs depends on the amount of voltage on the grid of the thyratron. If the grid is made less negative, a lower plate voltage will cause ionization. When ionization occurs earlier the amplitude of the waveshape may decrease on one hand, but on the other, since the discharge time is the same, the wave will repeat more often, that is, its frequency will become higher. Another result is that the charging curve is much more linear, that is, its slope is more constant.

Two facts of interest concerning thyratrons are that the de-ionization potential depends entirely on the gas pressure within the thyratron envelope and that de-ionization always occurs at the same potential regardless of the grid voltage or the applied plate potential. For example, the 884 tube, one customarily used in time base generator circuits, de-ionizes at about 20 volts.

The circuit which was described is called a free-running circuit; that is, so long as a d-c potential is applied, it continues to generate one time base after another. There are other circuits which are discussed later that must be driven. These generate only one time base and then wait for the driving voltage to start the charging process which generates the next time base.

DETERMINATION OF OPERATING FREQUENCY.—To calculate the frequency of oscillation of a thyratron sweep generator there are a number of factors that you must know. To become acquainted with these factors note figure 3–32 which shows a practical thyratron sweep-generator circuit. The applied voltage is 250 volts. The 50-K resistor in the plate circuit provides a minimum resistance in series with the tube. The 10-K resistor in the grid circuit limits the grid current to a safe value. To show how the operating frequency may be determined the following explanation is given. In this explanation it is assumed that the bias is set at minus eight volts and the variable plate-load resistor is set at 316K, as shown in figure 3–32. The curve in (B) shows the relation of the firing potential to the grid bias. Thus, when the grid voltage is minus eight volts, the tube ionizes when the plate-to-cathode voltage reaches 75 volts. Using this information, you can plot a charging curve like that shown in (C). This shows that during the first cycle the capacitor voltage starts from zero, but on the second and succeeding cycles it always starts charging from 20 volts, the de-ionization potential. At 75 volts, the thyratron ionizes, stopping the capacitor charge. With 230 volts applied, therefore, one cycle has a duration equal to the time required to charge the capacitor from 20 to 75 volts, or 55 volts.

To determine what fraction of full charge is represented by 55 volts, use the universal time constant chart curve. From it you learn that this value is 55/230, or 0.239. Therefore the capacitor is charged 23.9 percent before ionization. Using curve *A* in a universal time constant chart curve, find 0.239 of full charge, then move across to the curve, and down to 0.26 time constants. One time constant is *RC*, or 366K × 0.01, or 0.00366 second. Therefore, the time required for the charge is

0.26 × 0.00366, or 0.000952 second. Since frequency is the reciprocal of the time of one cycle, $F = 1/0.000952$, or 1,050 c.p.s.

To calculate frequency directly, use the following equations:

$$F = \frac{1}{2.302RC \log_{10} \dfrac{E_{bb} - E_d}{E_{bb} - E_i}}$$

Where,

F = Frequency in cycles per second.

RC = Time constant.

E_{bb} = D-c applied voltage.

E_i = Ionization potential.

E_d = De-ionization potential.

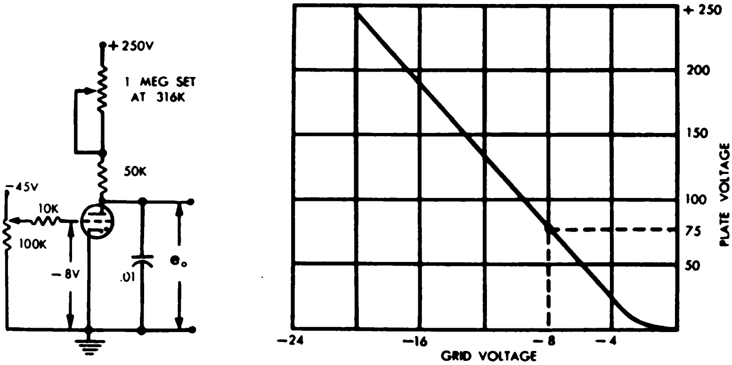
$$F = \frac{1}{2.302 \times 3.66 \times 10^5 \times 10^{-8} \log_{10} \dfrac{250 - 20}{250 - 75}}$$

$$F = \frac{1}{8.4275 \times 10^{-3} \log_{10} 1.31}$$

$$F = \frac{1}{8.48 \times 10^{-3} \times 0.1186}$$

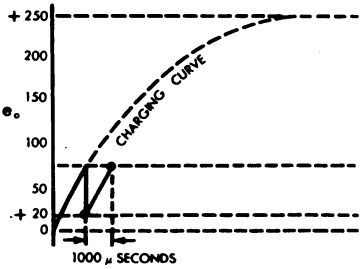
$$F = 1{,}000 \text{ c.p.s.}$$

EFFECT OF VARYING CIRCUIT COMPONENTS.—Any variation in the circuit constants affects frequency, linearity, and amplitude. Thyratron time based generators used in commercial oscilloscopes are required to produce output which can be varied in frequency. In these sets coarse frequency changes are provided by means of switching capacitors and fine frequency adjustment by means of varying the plate resistance and holding the grid bias constant. The usable frequency range extends from a few cycles to 50 kc. in the commercial oscilloscope. The upper frequency limit is 50 kc. because the discharge time at 50 kc. is equal to the charging time, and half of



(A) CIRCUIT

(B) IONIZATION CHARACTERISTICS FOR TYPICAL THYRATRON TUBE



(C) CHARGING CURVE

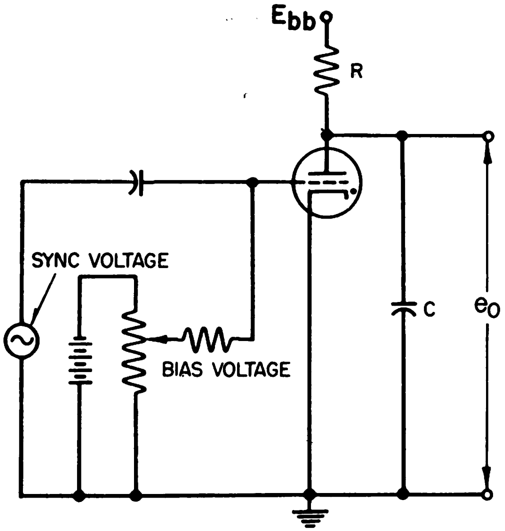
Figure 3–32.—Practical thyratron sweep-generator circuit.

any signal being investigated will be lost during the flyback (discharge) time.

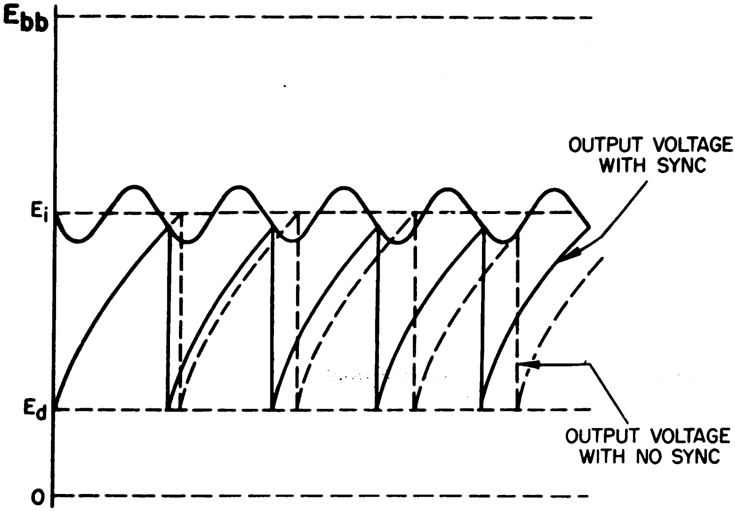
SYNCHRONIZATION.—In addition to all the factors just mentioned, the temperature of the gas, the slightly variable ionization potential, and the load on the circuit also affect frequency. Because of these factors, the circuit is usually not stable enough to maintain an adequately constant frequency by itself and must be synchronized with an a-c voltage that has a constant frequency. This synchronizing voltage may be any voltage that has a frequency near (usually a little above) the multiple, submultiple, or fundamental frequency of the thyratron generator. Due to the synchronization voltage, the thyratron tube ionizes slightly ahead of normal time, and the capacitor discharges immediately and starts a new cycle. Diagram (A) of figure 3–33 shows a typical synchronization circuit in which a sine wave is used to synchronize the generator in the grid circuit. The sine wave varies the grid voltage and this in turn varies the ionization potential.

The waveshape in (B) shows how the ionization potential varies at a slightly higher frequency than the natural free-running frequency of the generator. The dotted sawtooth waveshape results from a constant ionization (no synchronizing signal). At the time the capacitor voltage is about to reach the normal value, the ionization potential drops. This drop is due to a synchronizing signal on the grid. When the capacitor voltage reaches the ionization potential, the ionization occurs, starting a new cycle immediately. The same action recurs with each cycle and causes the time base frequency to "lock-in" step with synchronizing frequency. Often, in the use of the commercial oscilloscope, the waveshape being displayed on the CRT is used for synchronization. This is done so that the waveshape and time base will occur simultaneously, and the pattern will be stationary. If there is no synchronization, a difference in frequency will exist and the pattern displayed on the screen will move forward or backward along the time base.

TYPICAL APPLICATION OF THE GENERATOR.—A typical thyratron time base generator frequently used in oscilloscopes is shown in figure 3–34. The tube is a type 884 thyratron. Fixed bias is provided in the cathode by a voltage divider from B plus. The bias is held constant at 3.1 volts when the tube is nonconducting. Synchronizing voltage may be derived through a switch from one to three sources—the signal being displayed, the 60-cycle power line frequency or from a terminal post to which any other signal can be connected. The 100-K potenti-



(A) CIRCUIT



(B) WAVESHAPE

Figure 3–33.—Synchronization with sine wave.

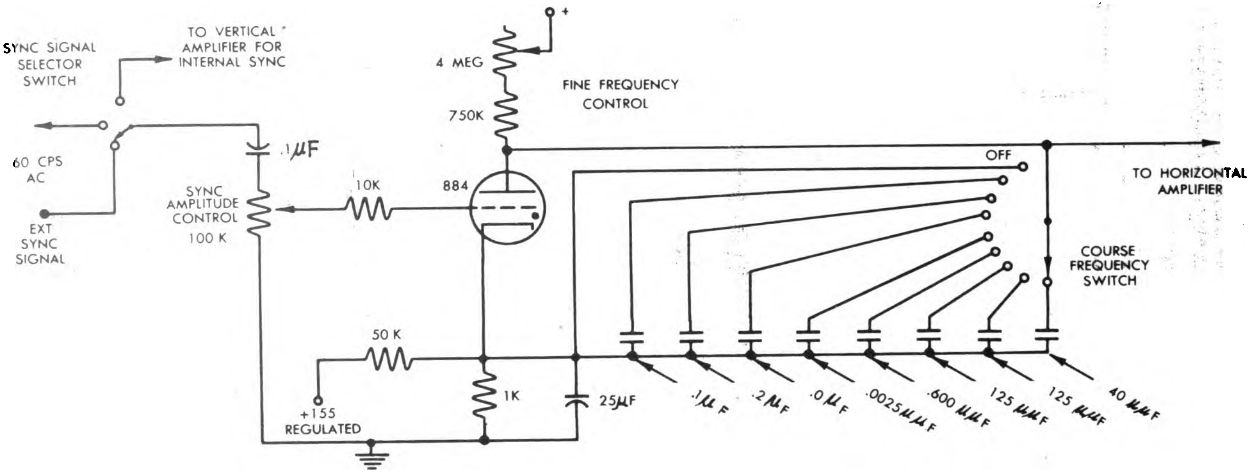


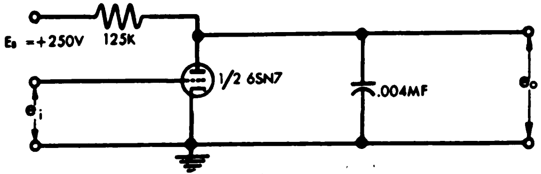
Figure 3–34.—A typical oscilloscope thyratron time base generator.

ometer provides for varying the amplitude of the synchronization voltage applied to the grid. One of the eight capacitors when connected in the R–C circuit provides for the coarse frequency adjustment. The 4-megohm potentiometer provides for fine tuning in that it can be used to set the frequency at exactly the desired value. Maximum linearity is obtained by keeping the amplitude of the output wave low. The desired amplitude is produced by amplifiers.

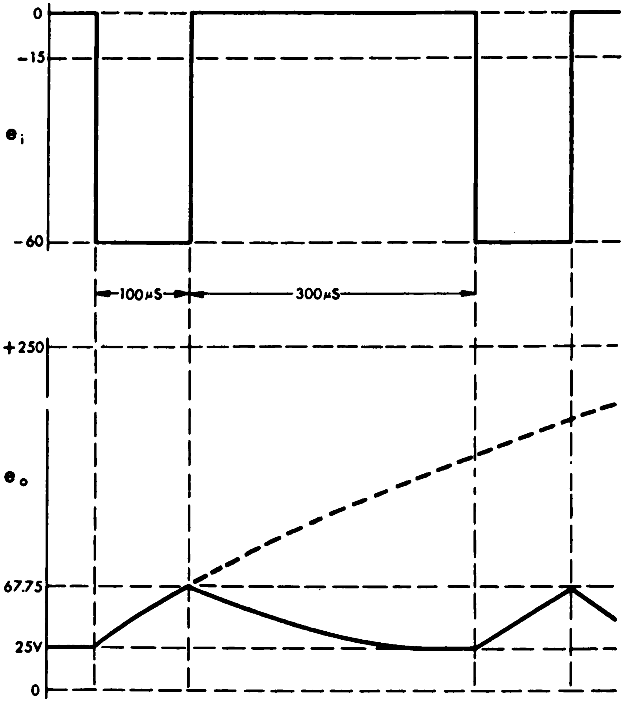
The Hard-Tube Sweep Generator

Because a thyratron is not readily adaptable for producing the time base with the spaced intervals required by radar and other electronic equipment, the high vacuum or hard tube is used instead. A hard tube is one that is highly evacuated; that is, one from which as much gas as possible is removed. In contrast, a SOFT tube is one with an incomplete vacuum; that is, one in which some gas remains in the envelope. A hard tube serves as a switch for discharging the capacitor but is not automatic and must be operated by an external circuit.

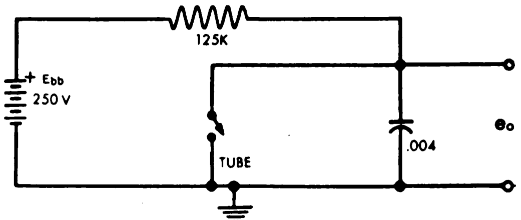
The circuit of a hard-tube sweep generator is shown in (A) of figure 3–35. In (B), the top waveshape shows the waveshape of the voltage that is usually used to operate the circuit. Normally, this voltage will hold the grid of the tube at zero. The tube in the circuit conducts a relatively large current. The loadline shown in (D) shows this current to be nearly 2 ma. According to the loadline chart, the plate voltage due to this current is 25 volts. In radar equipment a time base is formed whenever a radar pulse is transmitted. Simultaneously with the transmitted pulse, the negative square-wave input voltage drives the grid far below cutoff. As a result, the plate voltage tends to rise immediately to 250 volts, but since it cannot rise faster than the capacitor can charge, its rise is exponential. If a long time constant is used, the capacitor can only reach a small percentage of full charge before the grid cutoff voltage is removed, and tube current discharges the capacitor back



(A) CIRCUIT

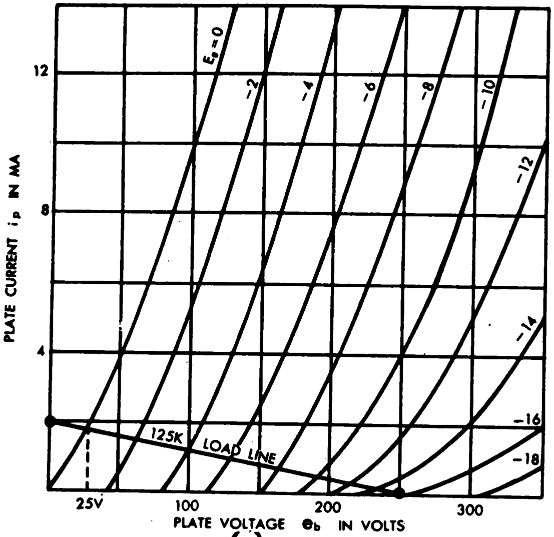


(B) WAVESHAPES

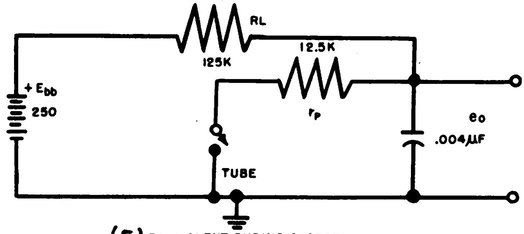


(C) EQUIVALENT DURING CHARGE

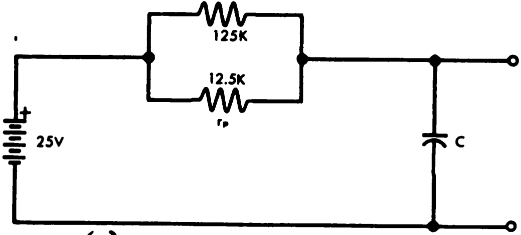
Figure 3–35.—Hard-tube sweep generator.



(D) LOAD LINE



(E) EQUIVALENT DURING CHARGE



(F) EQUIVALENT CIRCUIT DURING DISCHARGE

USING THEVENIN'S THEOREM

Figure 3–35.—Hard-tube sweep generator. Continued.

to 25 volts. The long time interval between transmitter pulses and the short time constant formed by the low tube resistance insure capacitor discharge to 25 volts before the next cutoff voltage occurs.

You can readily calculate the amplitude and linearity of a time base such as the one formed in the preceding discussion in the following manner. First, draw the loadline for the tube. The loadline shows that the voltage at the plate is 25 volts when the grid voltage is zero, and that cutoff for the tube is minus 16 volts when 250 volts is applied to the plate. Other factors to consider are that the minus 16-volt cutoff bias is easily exceeded by the minus 60 volts of the square pulse, which is 100 microseconds in duration (represented as time t in calculations), and that the applied voltage to the capacitor is 250 minus 25 or 225 volts. To find the capacitor voltage (e_c) after 100 microseconds, perform the following steps:

$$\text{time constant} = RC = 125 \times 10^3 \times 4 \times 10^{-9}$$

$$\text{time constant} = 500 \times 10^{-6}$$

$$\text{time constant} = 500\ \mu\text{sec.}$$

Since $t = 100$ microseconds, the capacitor is permitted to charge for 100/500 or 0.2 time constant. By referring to the universal time constant chart in chapter 2 you will find that after 0.2 time constant the capacitor will have reached a charge equal to 19 percent of the applied voltage. The capacitor voltage (C_c) equals the applied voltage (225 volts) times 19 percent plus the discharge voltage (25 volts) or

$$C_c = 225 \times 0.19 + 25 = 67.75 \text{ volts.}$$

The equivalent circuit during the charge is shown in (C) of figure 3–35. Here the tube acts as an open switch. If the rise continued at the initial rate, the curve would be perfectly linear. At the initial rate, the voltage would rise 225 volts during the first time constant. Since it is allowed to charge for 0.2 time constant, it would have risen 225 × 0.2 or 45 volts after this time. However,

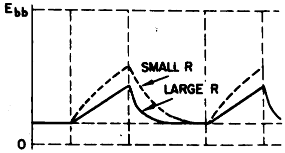
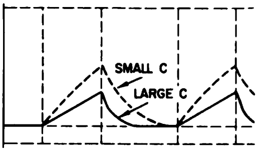
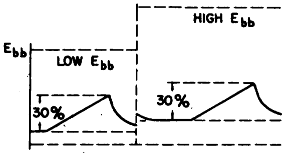
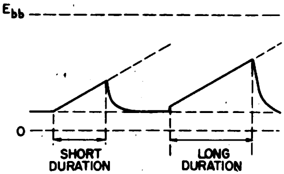
ELEMENT VARIED	EFFECT ON AMPLITUDE	EFFECT ON LINEARITY	EFFECT ON DURATION	WAVESHAPE
R IS INCREASED.	DECREASED. BECAUSE CONDENSER CHARGES SLOWER AND IS STOPPED BEFORE IT GETS AS HIGH.	IMPROVED. BECAUSE LOWER AMPLITUDE MEANS LESS OF CHARGING CURVE IS USED.	NO CHANGE. DURATION DETERMINED BY INPUT WAVESHAPE.	
C IS INCREASED	DECREASED. CAPACITOR CHARGES SLOWER AND IS STOPPED BEFORE IT GETS AS HIGH.	IMPROVED. LOWER AMPLITUDE MEANS LESS OF CHARGING CURVE IS USED	NO CHANGE DURATION DETERMINED BY INPUT WAVESHAPE.	
E_{bb} IS INCREASED.	INCREASED. CAPACITOR CHARGES TO SAME PERCENTAGE WHICH IS A HIGHER VOLTAGE.	NO CHANGE. CONDENSER CHARGES TO SAME PERCENTAGE AS BEFORE.	NO CHANGE.	
DURATION OF INPUT CUT-OFF VOLTAGE INCREASED.	INCREASED. CAPACITOR CHARGES LONGER AND SO TO A HIGHER VOLTAGE.	DECREASED. MORE OF CHARGING CURVE IS USED.	INCREASED. BECAUSE INPUT DURATION IS INCREASED.	
AMPLITUDE OF INPUT SIGNAL INCREASED.	NO CHANGE IF SIGNAL WAS ALREADY CUTTING TUBE OFF.	NO CHANGE IF SIGNAL WAS ALREADY CUTTING TUBE OFF.	NO CHANGE	NO CHANGE IN PLATE WAVESHAPE.

Figure 3–36.—Effect of varying elements in hard-tube sweep generator.

the actual rise is 42.75 volts (67.75 — 25). Thus, the actual curve varies from the linear curve by 2.25 volts (45 — 42.75).

When the grid voltage returns to zero, the equivalent circuit becomes as shown in (E) of figure 3–35 and also as the result of applying Thevenin's theorem as shown in (F). (This theorem states that any linear network containing one or more sources of voltage and having two terminals can be replaced by a simple generator having an internal impedance *A* and a generated voltage *E*, where *E* is the voltage that appears across the terminals when no load impedance is connected and *A* is the impedance across the terminals when all voltage sources are short-circuited.) The capacitor now discharges from 67.75 volts to 25 volts through the internal resistance of the conducting tube (R_p) and the load resistor (R_L). At zero grid voltage, the plate voltage is 25 volts and R_p equals 12.5K. The resistance of the tube in parallel with the 125-K load resistor makes a total resistance of 11.35K. The time constant of this R–C is 11.35 × 0.004 μf., or 45.5 microseconds. Because there are 300 microseconds between pulses, there is more than ample time for the capacitor to discharge to 25 volts.

The effect of varying elements in a hard-tube sweep generator is shown graphically in figure 3–36.

TYPICAL RADAR SWEEP GENERATOR.—As shown in figure 3–37, the typical sweep-generator circuit in a radar set has a negative square wave applied to the grid of the tube. The duration of the square wave is adjusted to exactly the time required for the radar wave to travel 10, 50, or 100 miles. When the range switch is on the 10-mile position, the input voltage is the 10-mile square wave, and the 0.003-μf. capacitor is switched in the plate circuit. For the other ranges, larger capacitors are switched into the plate circuit and the input square wave is longer. The 1-megohm potentiometers in the plate circuit vary the amplitude of the output sawtooth wave. They are adjusted so that the sweep line on the CRT screen will extend to the required length.

BOOTSTRAP SWEEP GENERATOR.—A disadvantage of the sweep-generator circuit in figure 3–37 is its inability to develop a linear sweep. This can be overcome by incorporating a bootstrap circuit.

The linear sweep-generator bootstrap circuit of figure 3–38 is essentially a circuit for generating a linear, saw-

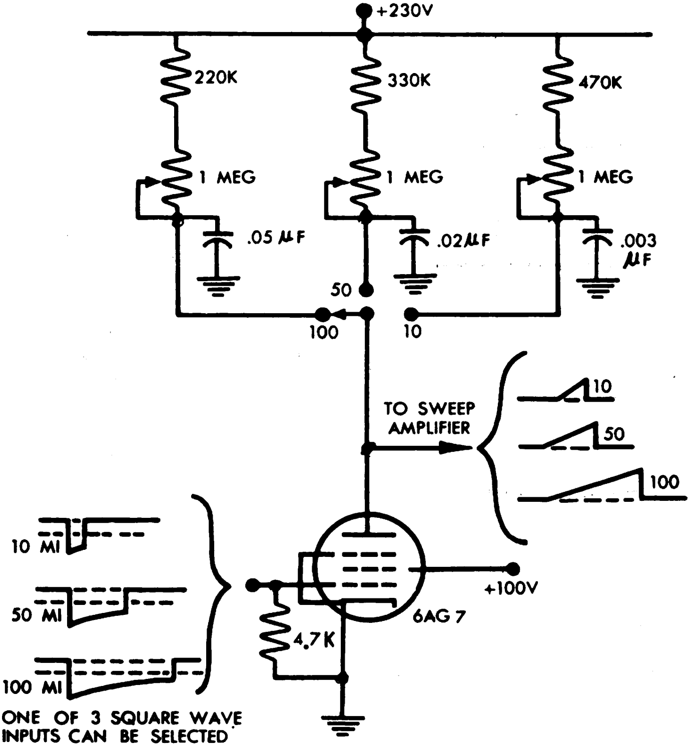


Figure 3–37.—Typical radar sweep generator.

tooth rise of voltage during the application of a negative rectangular pulse from the square-wave generator. To accomplish this, V_1 controls the charge and discharge of capacitor C_3. The cathode follower V_2 and the bootstrap diode V_3 provide a means of making the charging curve of the capacitor linear. In the absence of a negative gate from the square-wave generator both triodes V_1 and V_2 and the diode V_3 are conducting. Under these circumstances, most of the B-plus voltage is across R_3 and R_4, and capacitor C_2 is charged almost to the B-plus voltage. Capacitor C_3 is charged only to the low plate potential of V_1. When the negative gate input reaches the grid of V_1, it cuts off that tube. The timing capacitor in the plate

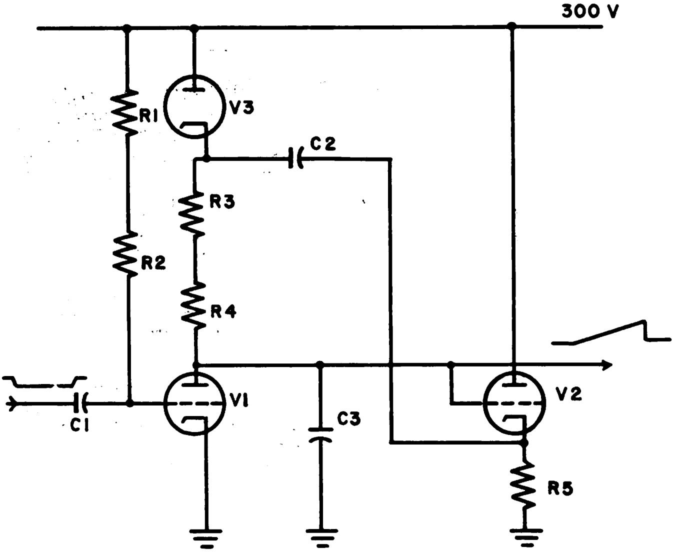


Figure 3–38.—A bootstrap sweep generator.

circuit of V_1 starts charging through R_3 and R_4. This rising voltage is in turn applied to the grid of the cathode follower V_2. The plate voltage of V_1 starts to rise toward the voltage of C_2, but through the cathode follower this same plate voltage rise is itself added to the voltage of C_2, so that the voltage drop across R_3 and R_4 remain constant as the plate voltage of V_1 rises.

Therefore, the current through R_3 and R_4 is constant, and voltage on the grid of V_2 rises linearly. The bootstrap diode V_3 ceases to conduct when the potential of its cathode rises above that of its plate, and effectively disconnects the charging circuit from the B-plus supply. Thus, the voltage of the linear waveform is not limited to the B-plus voltage.

Trapezoidal Waveshape Generator

The sawtooth sweep generator previously described is suitable for any electrostatically deflected CRT. However, you will find that the majority of the larger radar sets employ magnetic deflection instead in cathode-ray tube circuits in which sweep voltages are applied to coils. In long range sweeps, the sawtooth voltage is adequate for a linear sweep, but on short ranges, a special sweep voltage with a trapezoidal waveform must be applied to overcome the effects of the L–R formed by the sweep coils.

The deflection of the spot in a magnetic CRT depends upon a change in magnetic field strength. Since this magnetic field strength varies directly with the current in the deflection coils, any linear deflection of the spot requires a linear variation of the current through the deflection coil. A fairly linear current change occurs whenever a square wave is applied to an inductor if the first

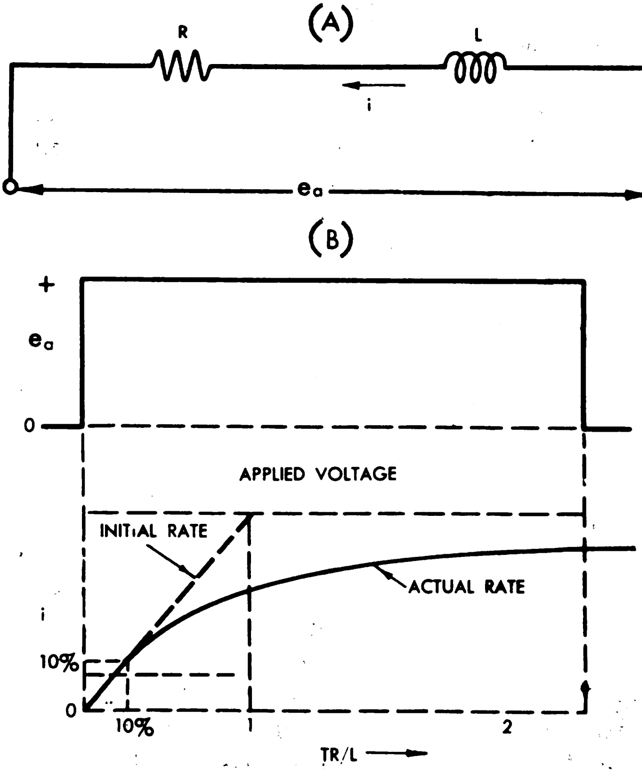


Figure 3–39.—Square wave added to L–R circuit.

10 percent of the time constant is used. As shown in figure 3–39, square wave e_a is applied to an L–R circuit. (Remember that all deflection coil circuits are L–R combinations because the coil wire and associated components introduce resistance in series with the actual inductance of the coil.) In this L–R circuit the current change through the coil increases rapidly at first but then decreases when the voltage drop across the resistor decreases the voltage applied to the coil. During the first 10 percent of the rise, the increase in current is quite constant.

Making the L/R value much longer than the time of the applied voltage produces a reasonable linear sweep. Referring to the simple circuit in figure 3–40 (B) utilizing a square-wave input, you see a tube that is normally cut off by fixed bias. When the sweep is ready to start, the grid voltage is brought up to zero with respect to the cathode by a positive square wave. Plate current flows, increasing at a rate determined by the inductance and resistance. This resistance is the plate resistance of the tube and the resistance of the coil. The time constant of the L–R circuit formed is long, and therefore the current rise is slow. The pulse on the grid ends before the current becomes very high, thus keeping the current increase in the linear portion of the curve. When the tube is cut off the magnetic field collapses, causing an oscillatory transient. The transients can be prevented either by a change in the circuit or made harmless by blanking the spot during this time.

It is difficult to keep the series resistance small because the plate resistance of the vacuum tube involved is never small. Therefore shorter time constants (larger R in ratio L/R) are accepted, and the applied voltage is changed to compensate for the large R. Referring to (B) of figure 3–39, notice that if there were no resistance in this circuit, the current would increase indefinitely at the initial rate along the broken curve. This is because the

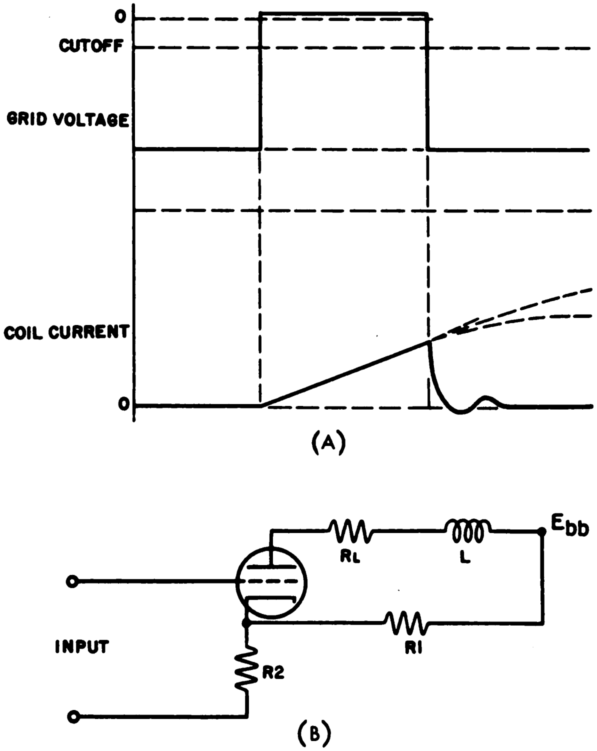


Figure 3–40.—Linear rise due to long time constant.

fundamental relationship between the voltage and current in an inductor is

$$E = L\ di/dt$$

where E is the applied voltage, L the inductance, and di/dt is the instantaneous rate of change of current with respect to time. L and E are constant and time changes linearly; therefore the current i must change linearly to make the equation true. The rate decreases along the exponential curve because the voltage drop across the resistor, due to this same current through it, subtracts from the applied voltage to leave less voltage across the inductor. Less voltage causes the current to increase at a slower rate.

The effect of the resistor may be eliminated if the applied voltage is increased at the same rate that the resistor voltage increases. Under these conditions, the subtraction of the resistor voltage maintains a constant voltage across the inductor, and the current increases at a constant rate.

ANALYSIS OF THE TRAPEZOIDAL WAVESHAPE.—The applied waveshape just described is called a TRAPEZOIDAL WAVE. In the waveshape shown in figure 3–41, the initial instantaneous rise is called the JUMP voltage. It is this voltage which starts the current flow in the inductor. The slow increase is called the SLOPE voltage. Its action is to compensate continuously for the drop across the resistor.

A quantitative analysis shows that there is a direct relation between the L/R, the jump voltage, and the slope voltage. In addition, since the trapezoidal voltage is produced with an R–C circuit, there is a direct relationship between the L/R of the sweep amplifier and the RC of the sweep generator.

The jump voltage is the voltage which would be applied to a pure inductance. It is a constant value equal to $L\,di/dt$, where L is the inductance and di/dt is the rate of current change in amperes per second. As the current change also occurs in the resistor, you can use the Ohm's law formula $E = IR$ to determine the voltage across the resistor. The facts are that R is constant and I is the same continuously changing current that is flowing through the inductor. Thus, according to Ohm's law the scope of the resistor voltage becomes $R\,di/dt$.

Important facts to remember are that in the determination of the waveshape required for any sweep coil, the ratio of the jump voltage to the slope voltage is the most essential quantity and that the jump to slope ratio is equal to the time constant of the L–R circuit. The relationship expressed mathematically reads,

$$\frac{\text{jump}}{\text{slope}} = \frac{L\,di/dt}{R\,di/dt} = \frac{L}{R}$$

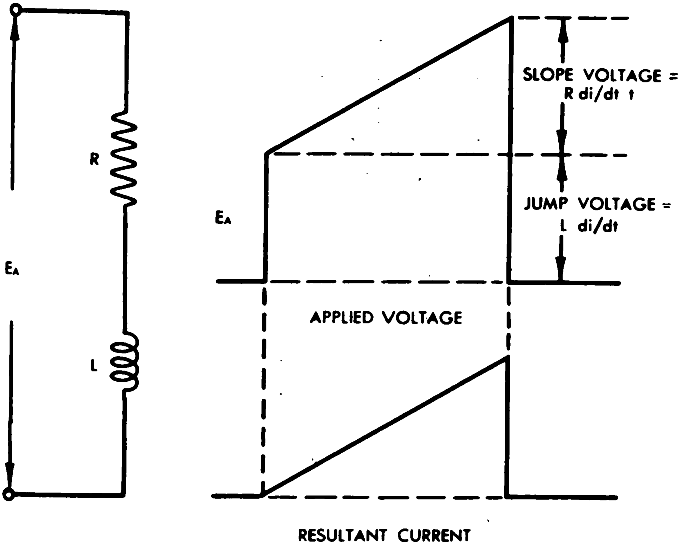
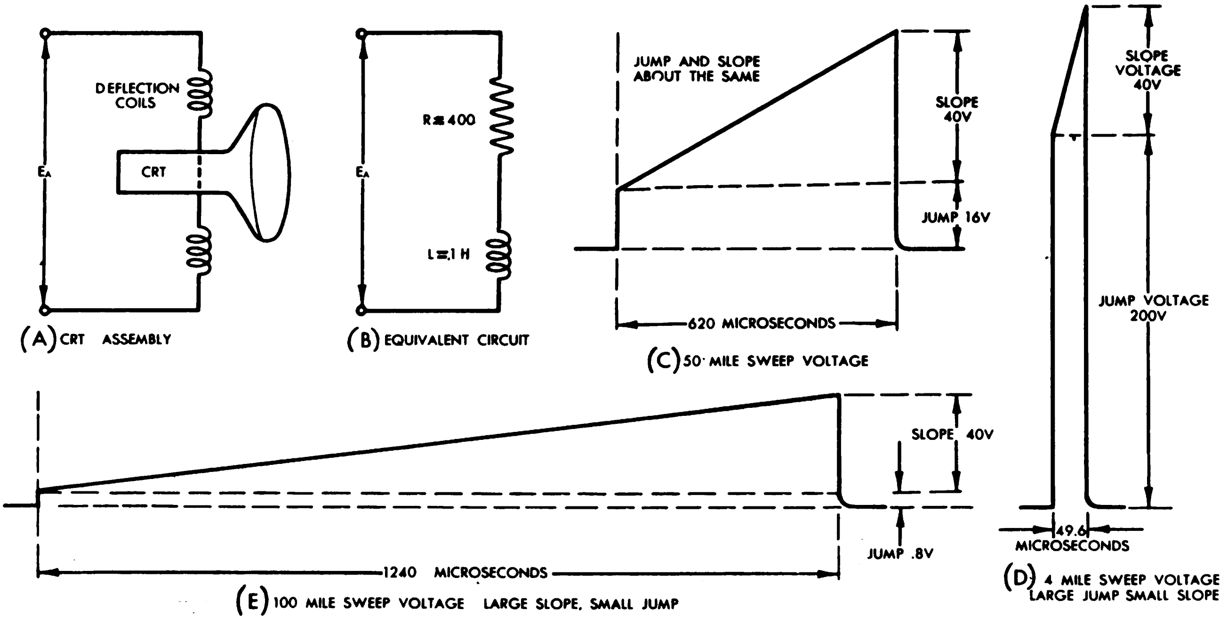


Figure 3–41.—Trapezoidal waveshape.

To learn to use this relationship, study the following sample problems:

Problem 1.—Assume the CRT and coils in the CRT assembly have been assembled and that as part of a radar set, a linear time base for a 50-nautical-mile range is required—that is, the spot must swing across the scope at a constant rate in 620 microseconds (the approximate time of 50 radar miles of travel). Experiments with d-c current indicate that a current of 100 ma. causes the spot to move the required distance. Therefore, for a linear time base that covers the screen, the current in the coils must change at a constant rate from zero to 100 ma. in 620 microseconds. The required current waveshape is shown in (C) of figure 3–42. Further measurements indicate that the total resistance in the coil windings is 400 ohms and the inductance is 100 millihenries. The equivalent circuit of the coils is shown in (B). What jump voltage and what slope voltage is required to cause a perfectly linear sweep on the CRT?



(A) CRT ASSEMBLY

(B) EQUIVALENT CIRCUIT

(C) 50 MILE SWEEP VOLTAGE

(D) 4 MILE SWEEP VOLTAGE LARGE JUMP SMALL SLOPE

(E) 100 MILE SWEEP VOLTAGE LARGE SLOPE, SMALL JUMP

Figure 3—42.—CRT assembly and waveshapes for 4-, 50-, and 100-mile ranges.

Solution:

The voltage across the inductor E_L is $L\ di/dt$.

Therefore,

$$E_L = \frac{10^{-1} \times 10^{-1}}{620 \times 10^{-6}}$$

$$E_L = 16.1 \text{ volts.}$$

The resistor voltage slope is $R\ di/dt$

Therefore, it equals $\dfrac{400 \times 10^{-1}}{620 \times 10^{-6}}$, or 6,450 volts per second.

The change in voltage across the resistor is at a rate of 6,450 volts per second. Since the time base only lasts 620 microseconds, the voltage change will be 620 millionths of this value or 6,450 × 620 × 10^{-6}, or 40 volts. You can easily check this result as follows: The current at the end of the time base is 100 ma. and is the final resistor current. The resistance is 400 ohms, therefore $E_R = IR = 0.1 \times 400$ or 40 volts. The required input waveshape, drawn to scale, is shown in (C) of figure 3–42.

Problem 2.—Notice the difference in the waveshape when the same sweep coil is used for longer and shorter sweeps. In each case the final current is 100 ma. for full scale deflection; therefore, the final resistor voltage is always 40 volts (100 ma. through 400 ohms). Find the jump voltage for a 4-mile range.

Solution:

The sweep time is 12.4 microseconds per mile. Therefore, for four miles it is four times that or 49.6 microseconds. Substituting in the equation $E_L = L\ di/dt$,

$$E_L = \frac{10^{-1} \times 10^{-1}}{40.6 \times 10^{-6}} \text{, or 200 volts.}$$

Since the 100 ma. is a smaller percentage of the final current when 200 volts is applied than when 16 volts is applied, the rise is more linear and the slope voltage is a smaller proportion of the total voltage. This condition approaches the applied square wave shown in figure 3–39. When the percentage of slope voltage becomes somewhat smaller, it may be safely omitted without serious non-linearity in current rise.

Problem 3.—Calculate the applied waveshape for the 100-mile range and find the final slope voltage when the final slope voltage is still 40 volts across the 400-ohm resistor and the time is 1,240 microseconds.

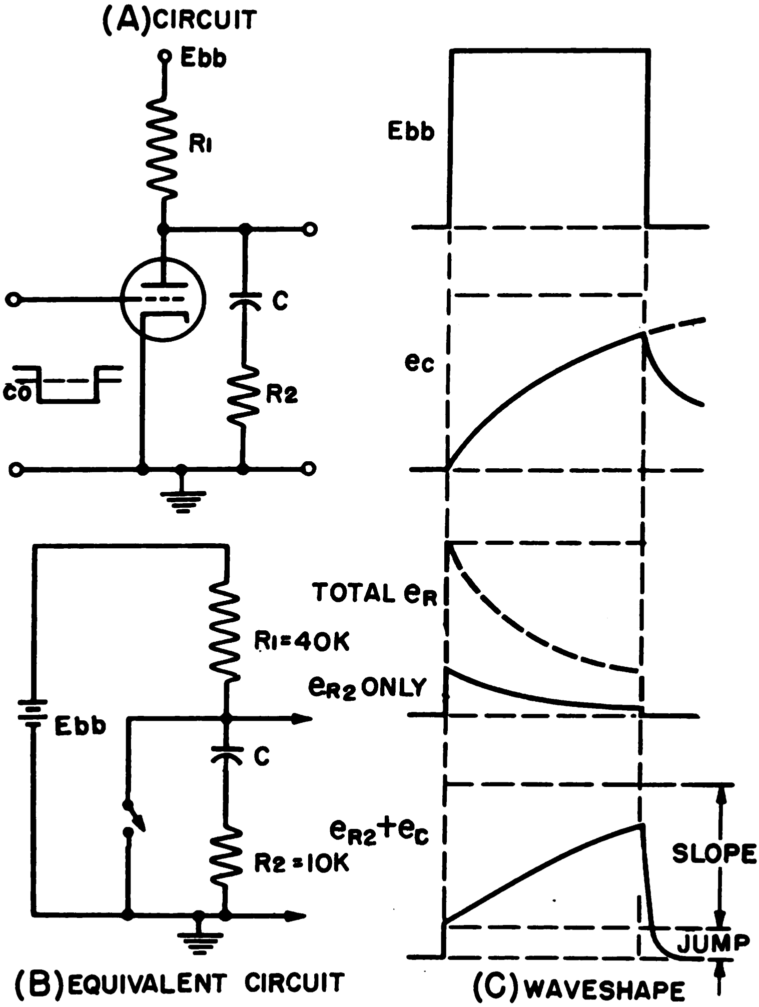
Solution:

Find the jump voltage, E_L, by the equation $E_L = L\,di/dt$

Substituting, $E_L = \dfrac{10^{-1} \times 10^{-1}}{1{,}240 \times 10^{-6}} = 0.8$ volt

Since the current rise is very slow, only a small voltage is applied to the inductor. In fact, the jump voltage is so small that it may be omitted altogether on this long range without appreciable change of the waveshape. The applied waveshape becomes a sawtooth waveshape similar to that used for the electrostatic CRT. The waveshapes for the 4-, 50-, and 100-mile ranges are drawn to scale in (C), (D), and (E) of figure 3–42. Notice that the small jump voltage goes with the long range and the large jump voltage goes with the very short ranges.

TRAPEZOIDAL SWEEP GENERATOR.—Where it is necessary to supply a trapezoidal waveshape to a sweep coil, a vacuum tube and R–C circuit may be used to produce the desired waveshape. Diagram (A) of figure 3–43 shows a simple vacuum tube circuit that will produce this kind of waveshape, and (B) shows the equivalent of the producing circuit. The triode serves only as a high-speed switch. When the tube is cut off by the driving pulse, the plate voltage rises toward the applied voltage at the rate the capacitor charges. However, some of the resistance of the R–C circuit is between the plate and ground. The high initial charging current flows through this resistor and the capacitor, raising the voltage immediately toward the applied voltage. In the


(A)CIRCUIT

(B)EQUIVALENT CIRCUIT

(C)WAVESHAPE

Figure 3–43.—Trapezoidal sweep generator.

curves, the capacitor voltage and total resistor voltage curve are shown in (C). If R_2 is one-fourth the size of R_1, one-fourth of the voltage will be across it at all times. The curve for the R_2 voltage is a miniature replica of the total resistor voltage. The output voltage is the sum of the R_2 voltage and the capacitor voltage. This total voltage is the trapezoidal voltage required for the sweep coil circuit.

As before, the two elements—jump and slope—form the waveshape in the sweep coil. The ratio of the jump to slope is L/R. To produce the same jump to slope ratio with an R–C circuit, it is necessary that the R–C circuit be designed so that R_2C is equal to the L/R of the sweep coil circuit. Mathematically, the relationship is expressed as $L/R = R_2C$.

This equation is derived as follows:

First, consider the jump voltage when the voltage E_{bb} is applied to the circuit. The total voltage E_{bb} appears across the resistors R_1 and R_2. The proportion of the voltage that appears across R_2 is expressed by the following equation:

$$E_{R2} = \frac{E_{bb}R_2}{R_1 + R_2}$$

If R_2 is much smaller than R_1, you can simplify the equation and get an approximate answer by dropping R_2 from the denominator. The equation now reads,

$$E_{R2} = \frac{E_{bb}R_2}{R_1}$$

Second, consider the slope, being sure not to confuse slope with the final slope voltage. Slope is a rate of so many volts per second while final slope is so many volts at the end of the rise. The initial charging rate of the capacitor is such that it would charge the capacitor to the applied voltage in one time constant. If only a small part at the start of the curve is used, then the initial rate is the slope. Since the voltage increases at a rate of one E_{bb} per time constant,

$$\text{Slope} = \frac{E_{bb}}{(R_1 + R_2)\ C}$$

When R_2 is quite small, you may disregard it for simplicity. Therefore, the slope equation reads,

$$\text{Slope} = \frac{E_{bb}}{R_1C}$$

Using the simplified equation, you can express the relation of the jump to slope as follows:

$$\frac{\text{jump}}{\text{slope}} = \frac{\dfrac{E_{bb}R_2}{R_1}}{\dfrac{E_{bb}}{R_1C}}$$

$$= \frac{E_{bb}R_2}{R_1} \times \frac{R_1C}{E_{bb}}$$

$$\frac{\text{jump}}{\text{slope}} = R_2C$$

You know this is an approximation, but you can see that the jump to slope ratio is nearly equal to the product of R_2 times C. For linear current increase, therefore, the time constant of the deflection coil circuit (L/R) should equal R_2C.

DESIGN CONSIDERATIONS IN A TRAPEZOIDAL GENERATOR.—In the practical utilization of the principles just described, it is usually necessary to make the sweep coil a part of the plate circuit of a power amplifier tube in order to obtain the high currents with the desired waveshape. The trapezoidal waveshape is then generated at a low level and amplified by the power amplifier tube.

To see how the details are worked out, study the example of a complete generator and sweep coil circuit in figure 3–44. The quantitative values worked out are for the coil used for producing the 50-mile sweep voltage described earlier. In the design of this circuit, you will find it convenient to start with the sweep coil and work backwards to the sweep generator. Therefore, the waveshapes which are shown and the analysis which follows are in reverse order.

The 50-mile sweep voltage coil has an inductance of 100 millihenries and a resistance of 400 ohms. A linear time base requires a linear current rise of 100 ma. in 620 microseconds, which can be accomplished with a

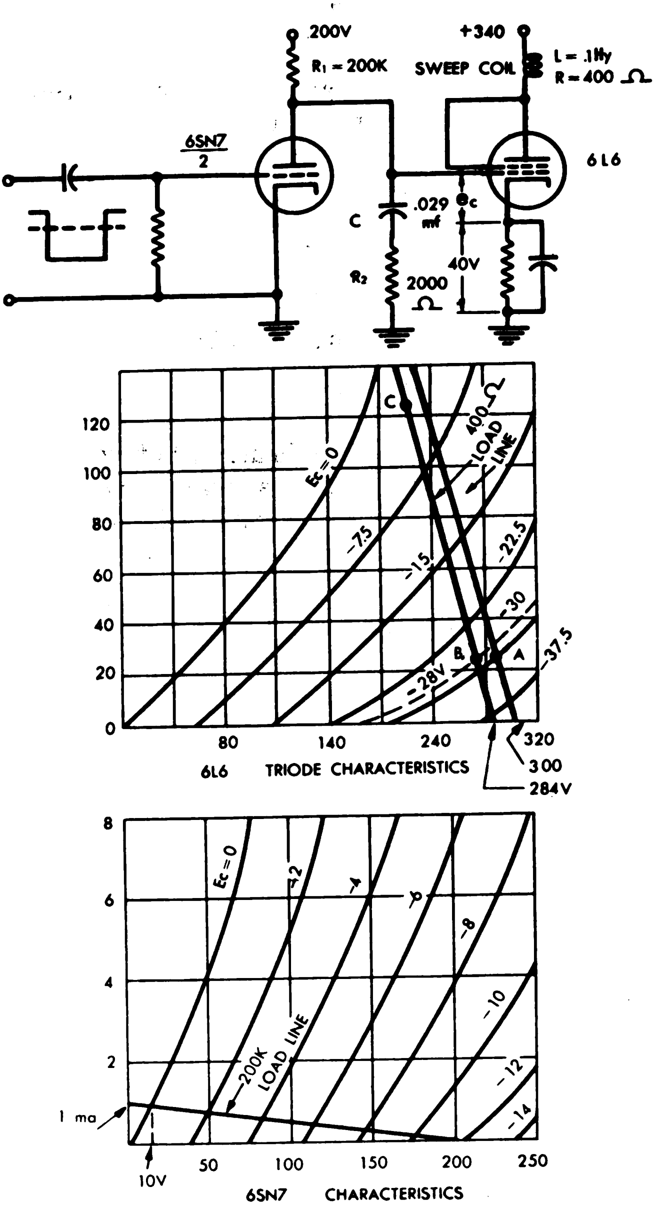
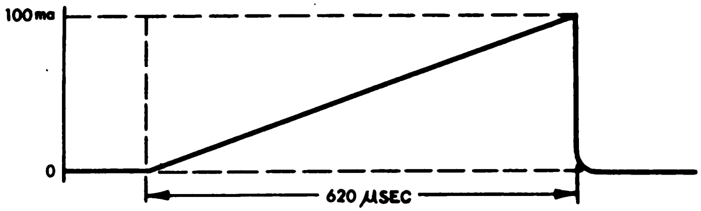
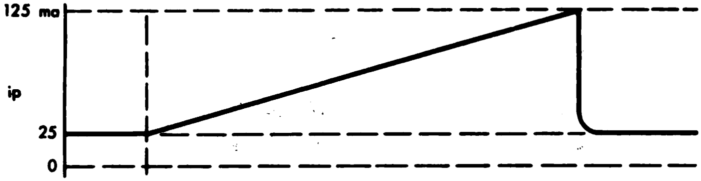


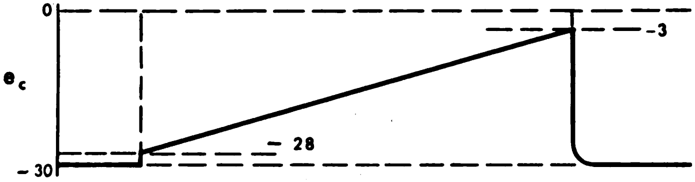
Figure 3–44.—Complete generator and sweep coil circuit.



DESIRED CURRENT CHANGE



CURRENT IN 6L6



GRID VOLTAGE ON 6L6

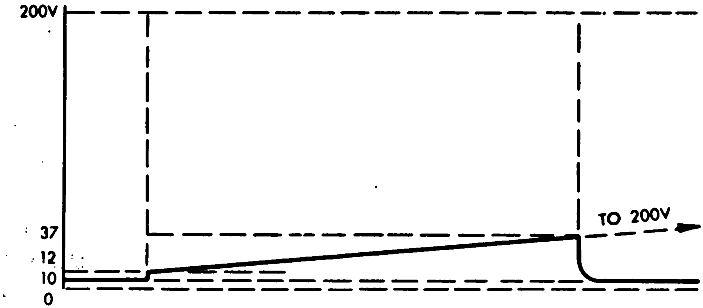
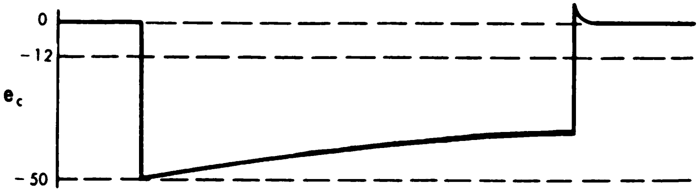


PLATE VOLTAGE ON GENERATOR TRIODE



GRID VOLTAGE ON TRIODE GENERATOR

Figure 3—44.—Complete generator and sweep coil circuit. Continued.

jump voltage of 16 volts (from previous calculations). Since this applied voltage is from the plate circuit and is due to grid voltage change, which is amplified in the tube, the required jump in grid voltage is only a fraction of the plate-voltage change.

In the circuit illustrated, the sweep coil is in the plate circuit of a 6L6 power amplifier. The 100 ma. current change actually starts from 25 ma. instead of zero to take advantage of the more linear part of the tube characteristics. The loadline is drawn from the 400-ohm resistance of the coil. The voltage on the grid between sweep voltages (first tube conducting) is plus 10 volts, but 40 volts bias makes the grid-to-cathode voltage minus 30 volts. Therefore, the waveshape starts forming at point *A* on the loadline. Then the sudden grid voltage change causes a jump of 16 volts in the plate circuit. The current increase in the coil generates 16 volts of back e.m.f., reducing the effective plate voltage by 16 volts. Thus a new loadline is drawn from 300 minus 16 or 284 volts. The grid voltage must change to point *B* on the loadline for a 16-volt change in the plate. The current has just started increasing above 25 ma., so point *B* is at 25 ma. plate current and 284 volts plate voltage. The grid voltage change is from minus 30 to minus 28 volts as read from these characteristic curves. Therefore, the jump required at the grid of the 6L6 is 2 volts.

After the jump, the slope occurs. The current must rise at a constant rate to 125 ma. At 125 ma. the grid voltage, as read from the characteristic curves, is minus 3 volts.

Note the difference in waveshape between the earlier circuit in (C) of figure 3–42 and the grid voltage curve for the 6L6 in figure 3–44. The slope component is many times as large as the jump component at the grid while the slope and jump are almost the same in the circuit in figure 3–42. The difference is due to the plate resistance of the 6L6 tube being in series with the coil. The additional resistance requires a much larger slope component.

The voltage change across the resistor R_2 and capacitor C in the circuit must be the same as the required grid voltage change. The sweep-generator components must be selected for this waveshape.

First consider the plate voltage required on the 6SN7. The slope must be linear, which means the charge must be limited to the first 10 percent of the charging curve. The voltage slope component has a final amplitude of 25 volts. The slope starts from 10 volts. If 100 volts are used as the applied voltage, 25 volts, being 25 percent of 100 volts, would be too nonlinear. Using 200 volts, and subtracting the 10 volts of original charge (plate voltage when the 6SN7 is conducting), 25 volts are 25/190 or 13.3 percent of full charge. This will be acceptable in this case because a higher voltage will raise the plate voltage, with the 6SN7 conducting, to more than 10 volts, which would require increasing of the 6L6 bias.

With 200 volts plate voltage, a load resistor must be selected which will hold the plate voltage at 10 volts while the 6SN7 is conducting and grid voltage is zero. As shown by the loadline on the 6SN7 characteristic curves, a 200-K resistor will accomplish this.

Next determine the capacitor size from the following data:

$$C_c = 25 \text{ volts}$$

$$E_{bb} = 200 - 10 = 190 \text{ volts}$$

$$t = 620 \text{ microseconds}$$

$$R = 200\text{K}$$

$$\text{The percent of full charge} = \frac{C_c}{E_{bb}} = \frac{25}{190} = 13.3 \text{ percent}$$

From the universal time constant chart in chapter 2, 13.3 percent of charge equals 0.107 time constant. This means that 620 microseconds is 0.107th of a time constant or 0.107 RC.

CIRCUIT CHANGE	EFFECT ON JUMP VOLTAGE	EFFECT ON SLOPE AMPLITUDE
INCREASE R_2	INCREASED BECAUSE RATIO R_2/R_1 IS INCREASED	NO CHANGE RC IS STILL APPROXIMATELY THE SAME
INCREASE C	NO CHANGE RATIO R_2 R_1 IS UNCHANGED	SLOPE IS LESS FINAL AMPLITUDE LESS BECAUSE RC IS LONGER, CHANGE IS SMALLER PERCENTAGE
INCREASE R_1	SMALLER RATIO R_2 R_1 IS LESS	GREATER BECAUSE RC IS SHORTER, C CHANGES TO HIGHER VOLTAGE
INCREASE E_{bb} (PLATE SUPPLY VOLTAGE)	INCREASED DUE TO GREATER APPLIED VOLTAGE	INCREASED DUE TO HIGHER VOLTAGE CHARGE LIMIT
DURATION OF DRIVING PULSE INCREASED	NO CHANGE SAME APPLIED VOLTAGE, SAME R_2 R_1 RATIO	NO CHANGE SAME APPLIED VOLTAGE, SAME RC_1

Figure 3–45.—Effect of varying elements in a trapezoidal sweep generator.

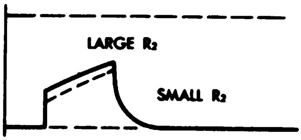
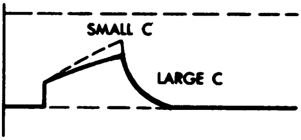
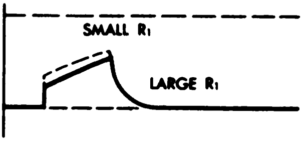
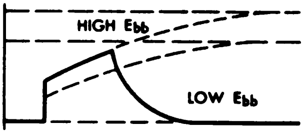
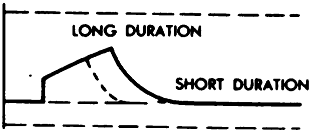
EFFECT ON SLOPE LINEARITY	EFFECT ON DURATION	WAVESHAPE
NO CHANGE RC IS UNCHARGED	NO CHANGE DURATION IS DETERMINED BY GRID PULSE	
IMPROVED BECAUSE SMALLER PART OF CHARGE CURVE IS USED	NO CHANGE	
DECREASED BECAUSE GREATER PART OF CHARGING CURVE USED	NO CHANGE	
NO CHANGE SAME RC SAME DURATION SAME PERCENTAGE	NO CHANGE	
DECREASED BECAUSE GREATER PART OF CHARGE CURVE USED	INCREASED BECAUSE INPUT PULSE IS LONGER	

Figure 3–45.—Effect of varying elements in a trapezoidal sweep generator. Continued.

Since $620 \times 10^{-6} = 0.107\ RC$
Then,

$$C = \frac{620 \times 10^{-6}}{0.107\ R} = \frac{620 \times 10^{-6}}{1.07 \times 10^{-1} \times 2 \times 10^{5}}, \text{ or}$$

$$C = 290 \times 10^{-10} = 0.029 \mu\text{f}.$$

The last item, the size of R_2, provides the jump voltage, which is 2 volts. You can calculate R_2 by transposing in the equation

$$\text{jump} = \frac{E_{bb} R_2}{R_1}$$

This gives,

$$R_2 = \frac{\text{jump} \times R_1}{E_{bb}}$$

Substituting,

$$R_2 = \frac{2 \times 2 \times 10^5}{2 \times 10^2} = 2{,}000 \text{ ohms}$$

The pulse required to cut off the 6SN7 according to the characteristic curves must be more than 12 volts negative. The negative square wave (shown as e_{c1}), having an amplitude of 50 volts, is more than ample for this purpose.

In the cathode-ray tube deflection coil circuit, a separate coil may be provided which will cancel the magnetic field caused by the residual 25 ma. current in the 6L6 tube. This causes the magnetic field strength to be zero between sweeps even though 25 ma. is flowing in the sweep coil.

In the chart in figure 3–45 which summarizes the effect of varying the sizes of the component parts of the trapezoidal sweep generator, notice that varying R_1, the plate-load resistor, affects the overall amplitude of the waveshape but has little effect on the jump-to-slope ratio. Varying the capacitor size changes the slope only, while varying R_2 changes the jump voltage only.

QUIZ

1. The function of a limiter, or clipper, is to restrict a signal in
   a. amplitude without affecting the shape
   b. amplitude without affecting the frequency
   c. frequency without affecting the shape
   d. frequency without affecting the amplitude

2. The plate resistance of a typical conducting diode is approximately
   a. 5 ohms
   b. 500 ohms
   c. 5,000 ohms
   d. 50,000 ohms

3. If a sine wave is applied to a positive series-diode limiter, limiting at zero potential, the output during the positive portion of the input will be
   a. a positive voltage
   b. a negative voltage
   c. zero volts
   d. both positive and negative voltage

4. In a negative series-diode limiter, the diode conducts
   a. when the input is positive
   b. when the input is negative
   c. at all times
   d. when the output is negative

5. The parallel-diode limiter differs from the series-diode limiter in
   a. having an output only when conducting
   b. having an output only when nonconducting
   c. being essentially a voltage-divider circuit
   d. limiting when biased below cutoff

6. In a negative shunt diode limiter, limiting at a negative potential, the output will be the same as the
   a. input
   b. positive portion of the input
   c. negative portion of the input
   d. input with the negative portion clipped at the limiting potential

7. A high amplitude sine wave is applied to a parallel-diode shunt limiter. Each limiter is limiting at a low bias voltage. The output will be a
   a. sine wave
   b. sine wave with positive portion clipped
   c. sine wave with negative portion clipped
   d. low amplitude square wave

8. In a grid limiter the operation of the grid circuit is the same as a
   a. positive limiter
   b. negative limiter
   c. positive clamper
   d. negative clamper

9. Which of the following statements about grid-limiting is true?
   a. Either positive or negative peaks may be clipped.
   b. Limiting occurs when electrons flow from the cathode to the plate.
   c. The amount of clipping depends on the ratio of grid-cathode resistance to series-grid resistance.
   d. A large resistor is required in series with the plate.

10. Saturation limiting with a small grid signal occurs in a tube in which the plate circuit contains a
   a. low plate-load resistor and a high plate-supply voltage
   b. high plate-load resistor and a high plate-supply voltage
   c. low plate-load resistor and a low plate-supply voltage
   d. high plate-load resistor and a low plate-supply voltage

11. An overdriven amplifier achieves both saturation and cutoff limiting by the use of a
   a. high plate-supply voltage, a high plate-load resistor, and a series-grid resistor
   b. low plate-supply voltage, a high plate-load resistor, and a large signal input
   c. high plate-supply voltage, a low plate-load resistor, and a small signal input
   d. low plate-supply voltage, a high plate-load resistor, and a series-grid resistor

12. In figure 3–17, the 8-μf. capacitor (in cathode of second triode) shorts. The plate waveform of V_2 would be
   a. a square wave
   b. negative spikes with switch in either position
   c. a low d-c voltage
   d. positive spikes with switch in either position

13. In figure 3–17, the 0.002-μf. coupling capacitor is increased to 0.2 μf. capacity. The waveform at plate of V_2 would be
   a. all positive spikes
   b. all negative spikes
   c. both positive and negative spikes
   d. high amplitude square wave

14. In radar indicators, clamping circuits are used to establish the d-c reference at the CRT because
   a. greater gain can be realized from use of many clamping circuits in series
   b. the clamping circuits synchronize the sweeps both horizontally and vertically
   c. the coupling capacitors between amplifier stages pass only the a-c components of a signal and reject the d-c reference level
   d. d-c coupling between amplifier stages cannot establish a d-c reference level

15. Clamping circuits are used in sweep circuits to
   a. limit the polarity of the sweep
   b. make the sweep linear
   c. limit the amplitude of the sweep
   d. start each sweep at the same potential

16. A negative-diode clamper will
   a. clamp peak of input to clamping potential
   b. clamp base of input to clamping potential
   c. limit the positive portion of input
   d. limit the negative portion of input

17. A positive-diode clamper will
   a. clamp peak of input to clamping potential
   b. clamp base of input to clamping potential
   c. limit positive portion of input
   d. limit negative portion of input

18. The action in the grid circuit of a grid clamper is the same as
   a. positive-diode clamper
   b. negative-diode clamper
   c. positive grid limiter
   d. negative grid limiter

19. In a thyratron sawtooth generator, increasing the bias voltage will
    a. increase sawtooth linearity
    b. increase frequency
    c. decrease amplitude of sawtooth
    d. decrease frequency

20. A thyratron oscillator can be stabilized in frequency by a synchronizing sinusoidal waveform only when the frequency of the synchronizing signal is
    a. exactly the same as the fundamental of the oscillator
    b. slightly below the fundamental of the oscillator
    c. slightly below the fundamental, harmonic, or subharmonic of the oscillator
    d. slightly above the fundamental, harmonic, or subharmonic of the oscillator

21. In a hard-tube sawtooth generator, increasing B+ will
    a. improve linearity of sawtooth
    b. increase amplitude of sawtooth
    c. increase frequency
    d. decrease amplitude of sawtooth

22. In figure 3–38, increasing the size of C_3 will
    a. increase amplitude of sawtooth
    b. decrease flyback time
    c. increase linearity of sawtooth
    d. decrease linearity of sawtooth

23. To obtain a trapezoidal waveform from a hard-tube sawtooth generator,
    a. a resistor is placed between sweep capacitor and ground
    b. a resistor is placed in parallel with sweep capacitor
    c. the sweep capacitor is replaced with a resistor
    d. two sweep capacitors are connected in series

24. In a radar set employing electromagnetic deflection, to switch from a long range to a short range, you should
    a. increase slope voltage only
    b. decrease slope voltage only
    c. increase both jump and slope voltage
    d. increase jump voltage only

CHAPTER 4

SPECIAL CIRCUITS—CONTINUED

SQUARE-WAVE GENERATOR

Some of the information dealing with multivibrators, as presented in this chapter, may seem repetitious of chapter 7, *Basic Electronics,* NavPers 10087. Since this chapter is written for Aviation Electronics Technicians First and Chief, the treatment of multivibrators must be more thorough. The preliminary data is given as an aid in introducing the more complicated aspects of the subject.

The multivibrator (square-wave generator) is a form of relaxation oscillator. There are several different types in use, depending on the function they are required to perform. They are used for time sharing, gating, counting, and timing bases. These circuits either produce a specified waveshape repeatedly as long as power is supplied or produced one wave when pulsed, and remain inactive until another pulse is applied to the circuit; then they produce another wave. The output waveshape is the same regardless of the waveshape or amplitude of the initiating or TRIGGER pulse. Those which run continuously with only power applied are called free-running multivibrators. Those which produce one complete cycle of operation for each synchronizing pulse are called a ONE-SHOT, ONE-CYCLE, or MONOSTABLE. Those which produce only a half cycle of operation for each synchronizing pulse are called BISTABLE, FLOPOVER or SCALE-OF-TWO circuits. The frequency of free-running multivibrators, as with other types of oscillators, is influenced by all voltages and component sizes associated with the circuits.

Therefore, although such circuits have only fair frequency stability when free running, adequate stability is readily achieved from a reliable a-c synchronizing voltage.

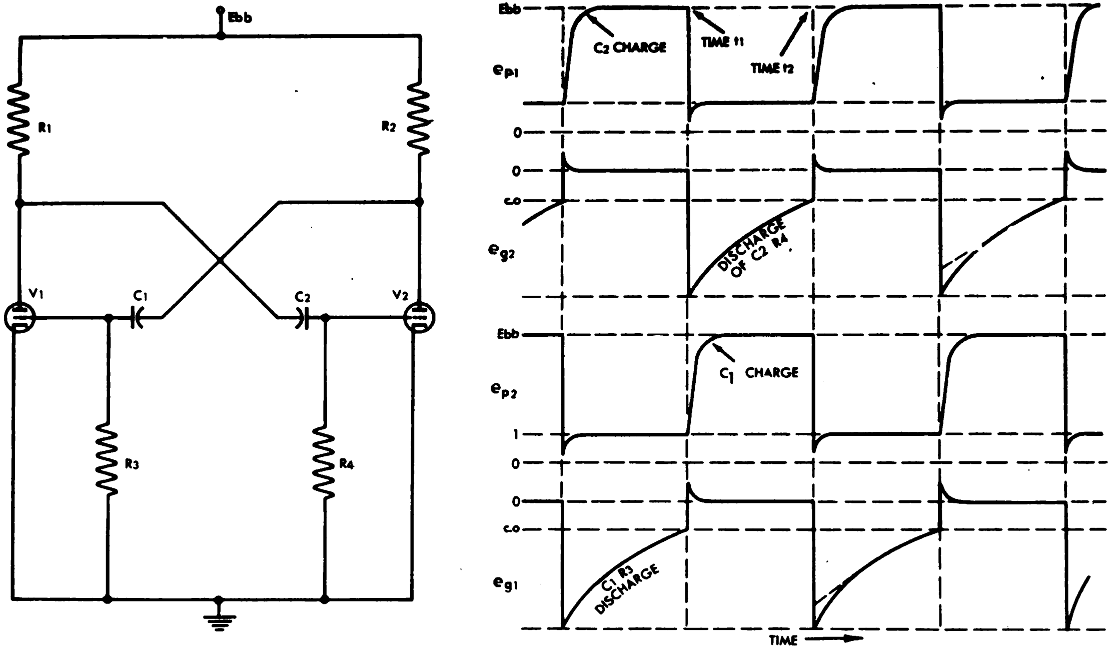
The square waveshape produced can be differentiated to produce accurately spaced peaked waveshapes. The duration and amplitude of any of the waveshapes are readily varied. Time delayed pulses can be produced by triggering a multivibrator with a peaked wave, and letting it go through a cycle and obtaining a new peaked wave somewhat later from the square wave produced.

The Plate-Coupled Multivibrator

The plate-coupled multivibrator is the simplest of the multivibrators. It produces a wave which is almost square when properly designed, and gives a perfect square wave when additions are made to the basic circuit. Note in figure 4–1 that basically this multivibrator consists of two triode amplifier stages, where the output of the second stage is introduced back into the input of the first. The phase inversion through the circuit is 360 degrees and any slight change in the first tube will be amplified through the circuit and will cause a greater change in the same direction in the first tube. That is, feedback is IN PHASE and the circuit will oscillate. There are no elements in the circuit to limit the feedback to any one frequency, so feedback occurs at all frequencies. Hence, a square wave containing several hundred frequencies is produced.

Looking at the action from the point of view of transients—that is, following each voltage change step by step—there are four different conditions that exist during one cycle of operation. They are as follows:

1. An extremely rapid change from V_1 conducting to V_2 conducting.
2. A long period during which V_1 is cut off and the circuit is relaxed.
3. A second violent change as V_1 conducts driving V_2 beyond cutoff.



BASIC MULTIVISRATOR CIRCUIT

WAVESHAPES

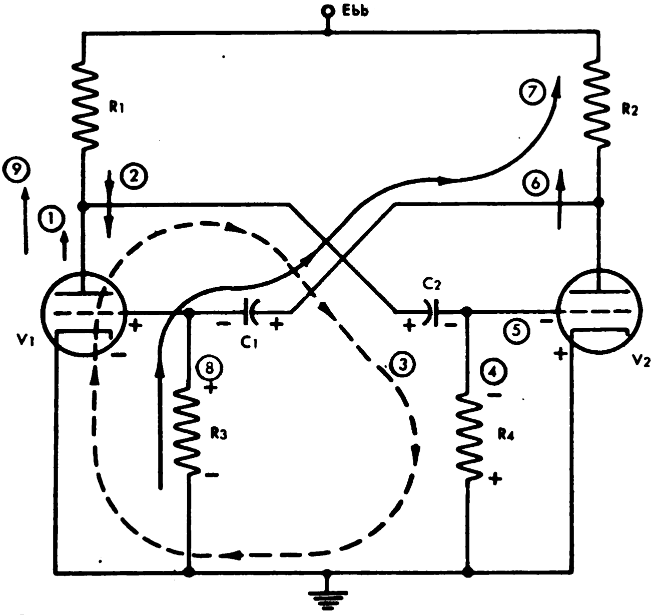
Figure 4–1.—Plate-coupled multivibrator.

4. A long period during which V_2 is cut off and the circuit is relaxed.

The cycle repeats as condition one follows condition four. The rapid changes just described are indicated by the vertical parts of the waveshapes in figure 4–1. These changes ordinarily occur in a fraction of a microsecond, and are so fast that the spot on a CRT becomes invisible when the waveshape is observed on an ordinary oscilloscope. The long relaxation periods are represented by the nearly constant voltages between switchover times.

CIRCUIT OPERATION.—When the filament voltages are on and the plate voltage is applied, current flows in both tubes. If the circuit components are the same in both tube circuits and both tubes are of the same type, the circuit is said to be SYMMETRICAL, and both currents should be equal. But it is impossible to obtain a perfect balance in the circuit, due to the nonuniformity of available parts and the irregular way in which emission occurs from the cathode. Any slight change in current in one tube that is not accompanied simultaneously by a similar change in the other tube will start oscillation.

Assuming that a slight current increase has occurred in the tube V_1, the following series of events then occur in rapid order: The increase of current in V_1 increases the voltage drop across R_1 which lowers the plate-to-ground voltage at V_1. This lower voltage is now less than the charge on C_2, and the capacitor C_2 discharges toward the new plate voltage. The discharge current flows through R_4 making the grid end negative. The negative grid of V_2 decreases the current on V_2. This reduces the drop across R_2, causing the plate-to-ground voltage at V_2 to be higher. Capacitor C_1 is between this plate and ground and therefore must charge to the new higher voltage. The charging current flows through R_3 toward the plate-supply voltage as shown by solid arrows in figure 4–2. This current causes the top of R_3 to be positive, and therefore the grid of V_1 to be positive. The positive grid causes the original slight increase of cur-



① CURRENT INCREASES IN V_1

② VOLTAGE DROPS AT PLATE OF V_1 DUE TO INCREASED VOLTAGE ACROSS R_1.

③ CAPACITOR C_2, NOW CHARGED HIGHER THAN APPLIED VOLTAGE, DISCHARGES TOWARD NEW LOWER VOLTAGE; DOTTED LINES SHOW CLOSED DISCHARGE CIRCUIT.

④ DISCHARGE CURRENT MAKES TOP OF R_4 NEGATIVE WITH RESPECT TO BOTTOM.

⑤ SINCE GRID AND CATHODE ARE CONNECTED TO R_4, GRID BECOMES NEGATIVE WITH RESPECT TO CATHODE.

⑥ MORE NEGATIVE GRID AT V_2 DECREASES CURRENT RAISING VOLTAGE AT PLATE OF V_2.

⑦ A CURRENT FLOWS THROUGH PATH OF ARROWS TO CHARGE CONDENSER C_1 TO NEW HIGHER VOLTAGE.

⑧ THE CHARGING CURRENT MAKES THE TOP OF R_3 POSITIVE WITH RESPECT TO BOTTOM AND CONSEQUENTLY, THE GRID OF V_1 IS POSITIVE.

⑨ THIS CAUSES A GREAT INCREASE IN CURRENT IN V_1, WHERE THE ORIGINAL SMALL INCREASES OCCURED.

Figure 4–2.—Sequence of changes in multivibrator.

rent to become a tremendous current increase. As the changes go around the circuit, a great voltage drop occurs at the plate of V_1 and grid V_2 is made more negative, current in V_2 is decreased more, plate voltage at V_2

goes up, making the grid of V_1 more positive and further increasing the V_1 plate current.

This is a cumulative process which continues until the current in V_2 is decreased to zero by a grid voltage far below cutoff. With zero current through V_2 the plate voltage becomes equal to the supply voltage since there is no voltage drop across R_2. C_1 quickly charges to this value by grid current flow in V_1. After that, current through R_3 is zero; hence the grid-to-cathode voltage is zero. The plate voltage of V_1 remains at a constant low value while the grid voltage is zero; so the capacitor C_2 continues to discharge to the value. While discharging, the current (dotted lines) holds V_2 cut off. Everything is at a standstill in the circuit except for the slow discharge of C_2. Of the four conditions listed at the start of this discussion on the plate-coupled multivibrator this corresponds to condition four. The slow discharge of C_2 continues and the voltage across C_2 becomes lower and lower. The voltage across R_4 continuously equals the C_2 voltage in accordance with Kirchhoff's voltage law, so the R_4 voltage decreases. Upon reaching a voltage just less than the cutoff voltage for V_2, a slight current starts flowing through V_2. This ends the period of inactivity in the circuit as swift changes follow this slight current.

The current lowers the V_2 plate voltage. C_1 starts to discharge, making the grid of V_1 negative. The resultant decrease in plate current raises the V_1 plate voltage. C_2 now recharges to the new V_1 plate voltage. The C_2 charging current makes the grid of V_2 positive and the original current is increased tremendously. This regenerative process continues until V_1 is cut off by the discharge of C_1. At this point, V_1 plate voltage rises quickly to the plate-supply voltage, C_2 charges swiftly to supply voltage through V_2 grid current, and the plate voltage of V_2 steadies at a low value. This ends the rapid changes during condition one.

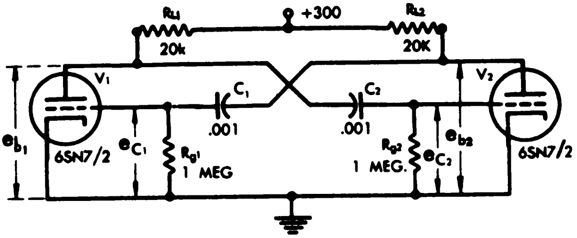
The second inactive period occurs as V_1 is held nonconducting by the slow discharge of C_1. The discharge continues until the C_1 voltage is just below cutoff for V_1. This period is as described for condition two.

Then the slight current in V_1 repeats the first rapid change described above. In the waveshapes in the basic circuit diagram in figure 4–1, the long period between time t_1 and t_2 is the time that V_2 is cut off, and the exponential grid-voltage decrease of e_{g2} shows how long the capacitor C_2 can hold V_2 nonconducting. At time t_2 the rapid change just described occurs, while the half cycle following that is the time V_1 is cut off.

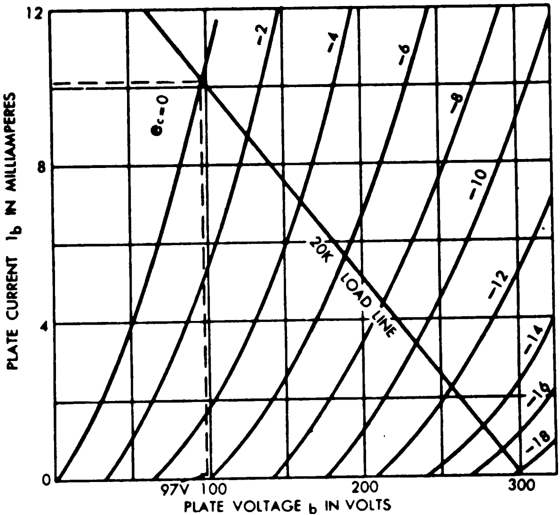
In summary, the operation of this circuit consists of long periods of time when one tube conducts a high current while the other tube is cut off, followed by an extremely rapid change to the other tube conducting and the first tube cut off.

The slow discharge of the capacitors is due to the long time constant of the discharge circuit. In the waveshape in the basic circuit diagram in figure 4–1, the grid voltage is shown as going positive for short periods of time. This occurs when the other plate is rising toward supply voltage. The capacitor is recharging, making the top of the grid resistor positive. The resultant positive voltage on the grid makes the grid act as a diode plate, and part of the emitted electrons flow to the grid and into the capacitor. This ready source of electrons charges the capacitor quickly. That is, the grid-to-cathode equivalent resistance is very small, forming a short *RC* during the capacitor charge. The positive-going grid causes an extra dip in plate voltage, which makes the grid of the other tube more negative than normal. The effect, if the grid did not become positive, is shown by dotted lines in the waveshapes in the basic circuit diagram.

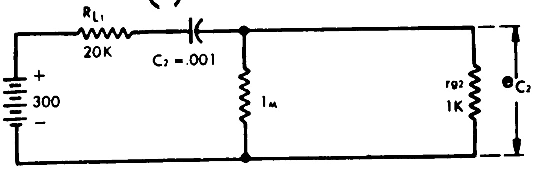
QUANTITATIVE ANALYSIS.—It is fairly easy to calculate the frequency and approximate voltage throughout a given circuit of this type. Consider, for example, the circuit shown in figure 4–3 (A). A plate voltage of 300 volts is applied to a 6SN7 dual triode with the circuit components shown. The plate voltage of V_1 varies be-



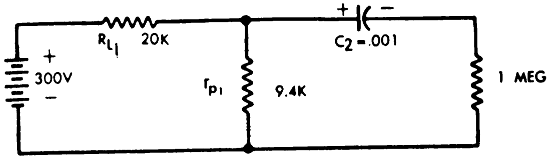
(A) CIRCUIT



(B) 6SN7 CHARACTERISTICS



(C) EQUIVALENT CIRCUIT FOR CHARGE OF C2



(D) EQUIVALENT CIRCUIT FOR DISCHARGE OF C2

Figure 4–3.—Quantitative analysis of plate-coupled multivibrator.

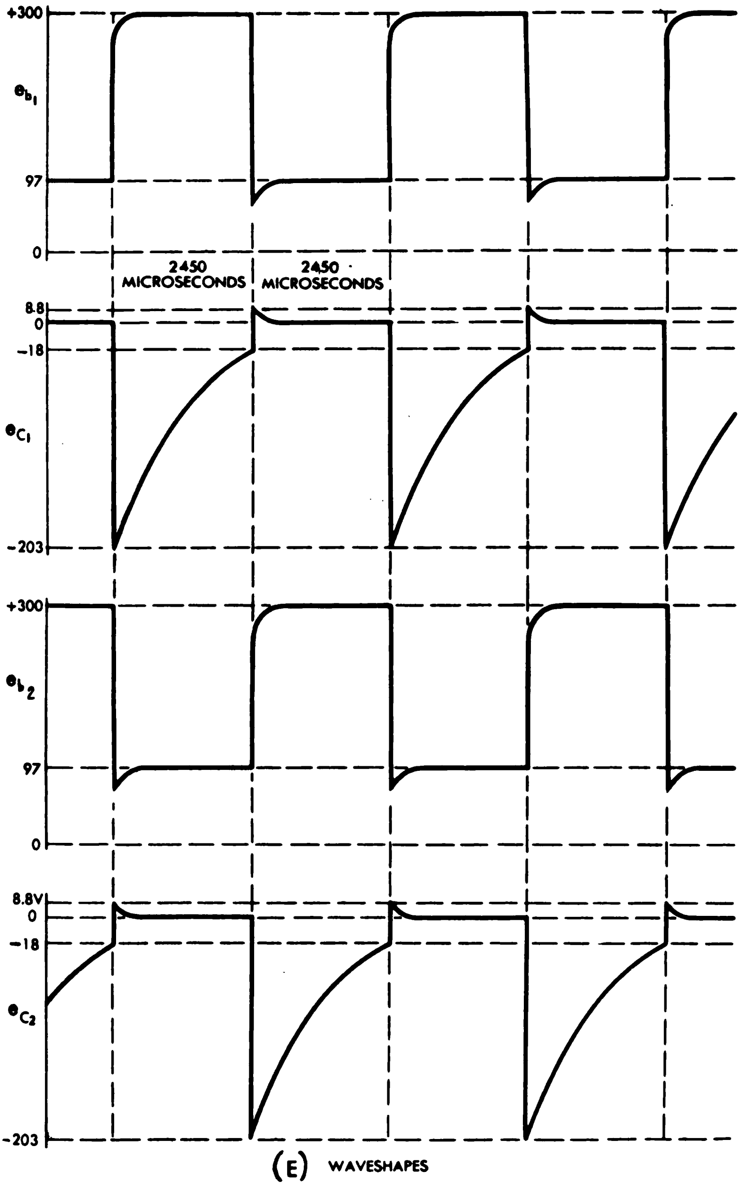


Figure 4–3.—Quantitative analysis of plate-coupled multivibrator. Continued.

tween the value corresponding to zero grid voltage and the value corresponding to zero plate current. You can determine these values from a 20-K loadline on the 6SN7 characteristic curves. The loadline crosses the zero grid-voltage line at a plate voltage of 97 volts. At zero plate current, the plate voltage is 300 volts. From this information you can construct the plate-voltage waveshape.

All this plate-voltage change occurs across the grid resistor on V_2 at the first instant. The resistor voltage is the grid voltage for V_2; so the grid will swing negative by the same amount that the plate voltage drops. The plate voltage drops 300—97, or 203 volts. The grid is at zero when the plate-voltage drop occurs, so the grid immediately goes 203 volts negative. Then the capacitor starts discharging down to 97 volts. The discharge will be exponential. The equivalent circuit during the discharge is shown in (D) of figure 4–3. The voltage across r_{p1} (the tube's plate resistance) is 97 volts and the capacitor is discharging from 300 volts to 97 volts. The plate resistance r_{p1} is found by using the plate current and voltage when the grid voltage of V_1 is zero.

$$r_{p1} = e_b/i_b = 97/10.3 \text{ ma.} = \text{about } 9.4\text{K ohms.}$$

In parallel with this (through the battery) is the 20-K plate-load resistor. This parallel combination will have a resistance of about 6K and will be in series with the 1-megohm resistor. For simplicity this 6-K resistance can be disregarded. This will only introduce an error of less than one percent. The discharge time constant without the 6-K resistance will be 0.001 μf.$\times$1 meg.=1,000 miscroseconds.

The discharge continues until the resistor (and grid) voltage at V_2 is just above cutoff. Cutoff voltage for this tube with 300 volts applied is minus 18 volts as read from the characteristic curves. The cutoff period is terminated when the voltage drops to 18/203 or 8.86 percent. The resistor voltage is down to 8.86 percent in 2.45 time constants. This is calculated from the capacitor discharge curve on the universal time constant chart in chapter 2. Since the discharge time constant is 1,000 microseconds, the cutoff time equals 2.45 time constants multiplied by 1,000 microseconds or 2,450 microseconds.

The cutoff time for tube V_2 will be 2,450 microseconds. Since the resistor and capacitor values are exactly the same, tube V_1 will be cut off for 2,450 microseconds during the other half cycle of operation. The grid and plate-voltage values and curves will also be identical. If a difference exists in the circuit of each tube, separate calculations must be carried out for each half cycle. In this case, $2{,}450+2{,}450=4{,}900$ microseconds. The frequency at which square waves are produced is the reciprocal of the duration.

$$F=\frac{1}{4{,}900}\times 10^{-6}=204 \text{ cycles per second.}$$

The duration and amplitude of the positive peak on the grid waveform can be readily calculated from the equivalent circuit shown in (C) of figure 4–3. The positive peak occurs as the capacitor is charged to a higher plate voltage. The capacitor charges through R consisting of the plate-load resistor for V_1 and the grid-cathode resistance of V_2. This is a total of 21K. The 1-megohm resistor carries only 0.1 percent of the charging current and is therefore safely disregarded. With the 0.001-μf. capacitor the time constant is $0.01\times 10^{-6}\times 2.1\times 10^{-4}$, or 21 microseconds. If the charge requires 5 time constants, the positive peak will disappear after a hundred microseconds. The amplitude of the peak can be determined by using Ohm's law. The total voltage rise at the plate of V_1 is 203 volts. This raises the grid voltage from cutoff of minus 18 volts to some positive value. Since the first 18 volts of plate-voltage rise change the grid voltage to zero, the remaining rise above zero is 203 — 18, or 185 volts. This causes grid current to flow and the voltage to divide across the series combination consisting of the grid-cathode resistance and the plate-load resistance. As the grid-cathode resistance equals 1,000 ohms,

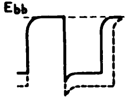
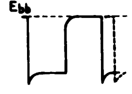
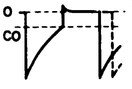
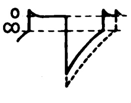
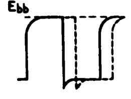
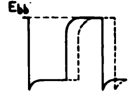
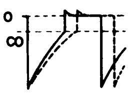
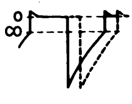
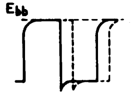
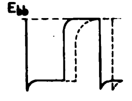
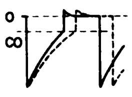
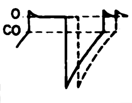
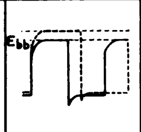
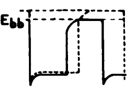
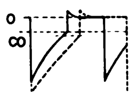
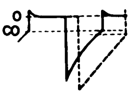
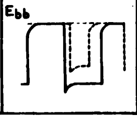
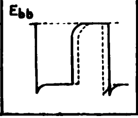
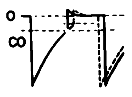
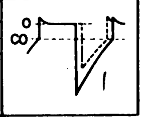
CHANGE	FREQUENCY	AMPLITUDE OF e_{b1}	DURATION e_{b1} AT E_{bb}	AMPLITUDE e_{b2}	DURATION e_{b2} AT E_b	WAVESHAPE CHANGE ——— BEFORE CHANGE -------AFTER INCREASE			
						e_{b1}	e_{b2}	e_{g1}	e_{g2}
$R_{l,1}$ INCREASED	DECREASED. V_2 CUT OFF LONGER.	INCREASED. e_{b1} IS LOWER DURING ZERO GRID VOLTAGE TIMES	UNCHANGED	UNCHANGED	LONGER. GRID IS MORE NEGATIVE AT START OF C.O TIME				
R_{g1} INCREASED	LOWER. V_1 CUTOFF LONGER	UNCHANGED	LONGER. TUBE IS CUT OFF LONGER	UNCHANGED	UNCHANGED				
C_1 INCREASED	LOWER. V_1 CUTOFF LONGER	UNCHANGED	LONGER	UNCHANGED	UNCHANGED				
E_{bb} INCREASED	LOWER. BOTH GRIDS DRIVEN BELOW CUTOFF. CUT OFF VALUE LOWERS SLIGHTLY.	INCREASED. E_b GOES UP. e_{b1} VALUE FOR ZERO GRID VOLTAGE GOES UP SLIGHTLY.	INCREASED. V_1 IS CUTOFF LONGER.	INCREASED. E_b GOES UP A LOT. e_{b2} VALUE FOR ZERO GRID VOLTAGE GOES UP SLIGHTLY.	INCREASED. V_2 CUTOFF LONGER.				
V_1 REPLACED WITH TUBE OF HIGHER Mu.	HIGHER	DECREASED. HIGH r_p OF V_1 CAUSED e_{bb} FOR ZERO GRID VOLTAGE TO BE HIGHER.	SLIGHTLY LONGER. CUTOFF FOR V_1 IS LOWER.	UNCHANGED	DECREASED SINCE GRID OF V_2 IS NOT DRIVEN VERY FAR NEGATIVE.				

Figure 4—4.—Effects of varying elements in multivibrator.

and the plate-load resistance equals 20,000 ohms, $\frac{1}{21}$ of the 185 volts, or 8.8 volts appear across the grid-cathode resistance. Thus the positive peaks at the grid will be 8.8 volts above zero. Because the plate voltage cannot rise faster than the capacitor can charge, the plate-voltage rise is exponential over a period of 100 microseconds.

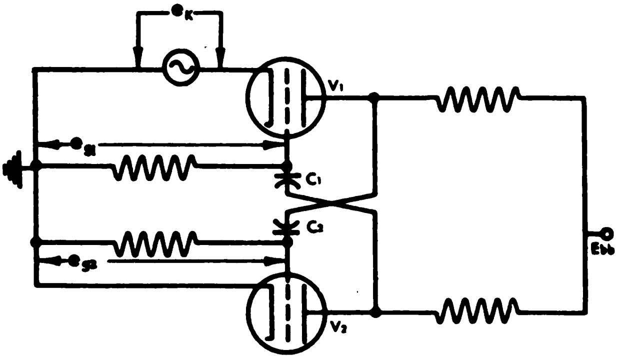
This rounds off the leading edge of the plate waveshape. The positive grid at the other tube causes a dip in plate voltage at that tube. The output from this circuit is normally taken from either plate although the grid voltage is sometimes useful.

In the chart (fig. 4–4) showing the effects of varying the sizes of the component parts of this circuit, the essential data is given, but the detailed and less important effects are omitted because of a lack of space. The effects of changing R_{L2}, R_{g2}, C_2, and V_2 are not listed since these effects are similar to those that result from changing R_{L1}, R_{g1}, C_1, and V_1 respectively—except that each effect occurs with the other half cycle of operation.

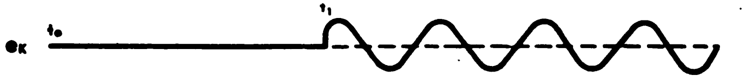
Synchronization of the Multivibrator

Free-running multivibrators are generally not used as such in electronic circuits because their frequency stability is poor. To avoid this frequency instability, multivibrators are usually synchronized with another frequency which forces the period of the multivibrator oscillation to be exactly the same as the period of the synchronizing frequency. Such a multivibrator is said to be locked-in by the synchronizing voltage.

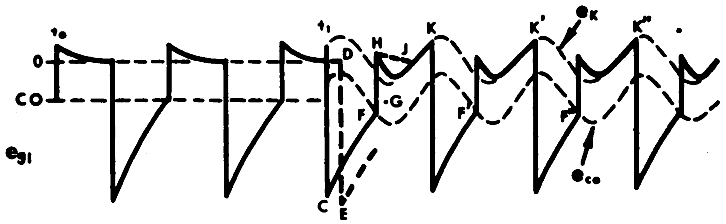
SYNCHRONIZATION WITH A SINE WAVE.—Although a waveform of almost any shape may be used for synchronization, a sine wave or a pulse is generally used. The circuit of a multivibrator which is synchronized by the injection of a sine-wave voltage in the cathode circuit of one tube is shown in figure 4–5 (A). The actual grid-to-cathode voltage of V_1, which is the voltage controlling the flow of plate current, is the difference between the grid-to-ground voltage e_{g1} and the cathode-to-ground voltage e_k. The source of sinusoidal voltage should have a



(A) SYNCHRONIZATION AT CATHODE



(B) CATHODE VOLTAGE



(C) GRID WAVESHAPE

Figure 4–5.—Synchronization with sine wave on cathode.

low internal impedance, otherwise the plate current of V_1 flowing through this source causes a voltage drop which alters the sinusoidal shape of the wave. Sometimes a low-voltage winding on a power-supply transformer, such as a filament winding, is used to supply this control voltage.

If the multivibrator in (A) of figure 4–5 were balanced and running freely, the waveform of the voltage at the grid of V_1 would be that shown between times t_0 and t_1 in (C). Because there is no synchronizing voltage applied between these, two times, the cathode voltage will be constant at ground potential as shown in (B). At time t_1, when the synchronizing voltage is suddenly applied, the cathode potential starts to vary sinusoidally. Although this variation does not effect the voltage on the grid with respect to ground, the grid-to-cathode potential contains this sinusoidal component of voltage. Therefore, the effective cutoff voltage of the tube varies sinusoidally about the normal value in phase with the synchronizing voltage on the cathode. The cathode voltage e_k and the effective cutoff voltage e_{co}, shown at (C), explain the synchronizing action by showing that V_1 conducts when the e_{g1} curve crosses the e_{co} curve.

At time t_1 the voltage of the cathode starts to rise, decreasing the conduction of V_1. The positive-going voltage produced at the plate of V_1 initiates the switching action, and the tube is quickly cut off. Thus e_{g1} drops along t_1 instead of *DE* as it would in the free-running condition. Capacitor C_1 discharges exponentially along curve *CFG*. Since this curve intersects the e_{co} curve at *F*, the switching action, by which V_1 is made conducting and V_2 is cut off, takes place at this time instead of at a short time later. Thus the switching takes place at time *F* instead of at *G* as would be the case in a free-running multivibrator. The switching drives the grid of V_1 positive, but the grid current drawn, quickly charges C_1 so that the grid returns to cathode potential. If e_{g1} followed curve *HJ* as it would if the multivibrator were free running, the grid would draw current longer, but the synchronizing voltage causes the cathode to be negative relative to ground at this time. Therefore, e_{g1} follows the cathode voltage along curve *HK*. When the cathode voltage begins to rise, the plate current through V_1 starts to decrease. By the time *K*, the rise in voltage at the plate of V_1 resulting from this decrease in plate current has become large enough to drive V_2 into conduction, and the tubes are very rapidly switched.

The cycle of the multivibrator is forced to be somewhat shorter than the length of the free-running cycle by the action of the sine wave. Thus switching in one direction occurs at points F, F', F'', and so forth, and switching in the other direction occurs at K, K', K'', and so forth. The period of the multivibrator, KK' or $F' F''$, is seen to be equal to the period of the sine wave after the short transition interval, therefore, the multivibrator may be said to be synchronized. A sine wave can thus be used to control the frequency of a multivibrator. The synchronizing voltage can make the multivibrator operate either above or below its natural frequency. However, if an attempt is made to pull the multivibrator to higher and higher frequencies, a limit is reached beyond which the multivibrator synchronizes to one-half of the driving frequency. Similarly, the multivibrator may synchronize to one-third or a smaller fraction of the driving frequency and **FREQUENCY DIVISION** may be obtained.

When the sine-wave synchronizing voltage is applied to the grid instead of the cathode of one of the tubes of the multivibrator, (fig. 4–6 (A)), the synchronizing voltage as shown adds directly to the grid voltage. Consequently, the grid-voltage waveform does not resemble that of a free-running multivibrator. Without the synchronizing signal, the multivibrator is free running and the grid waveform appears as shown between t_0 and t_1 in figure 4–6 (B). The synchronizing voltage is applied slightly before time t_1 in the random phase shown, but because of grid limiting, it has no effect until V_1 cuts off at its natural time t_1. Between C and D the synchronizing voltage adds to the normal discharge voltage which would rise along curve EF, and result in the distorted curve shown. Since the synchronizing frequency is higher that the natural frequency of the multivibrator, the distorted curve reaches cutoff sooner than curve EF.

When V_1 is conducting during the time between G and H, grid limiting occurs and prevents the synchronizing voltage from adding to the grid voltage. However, when

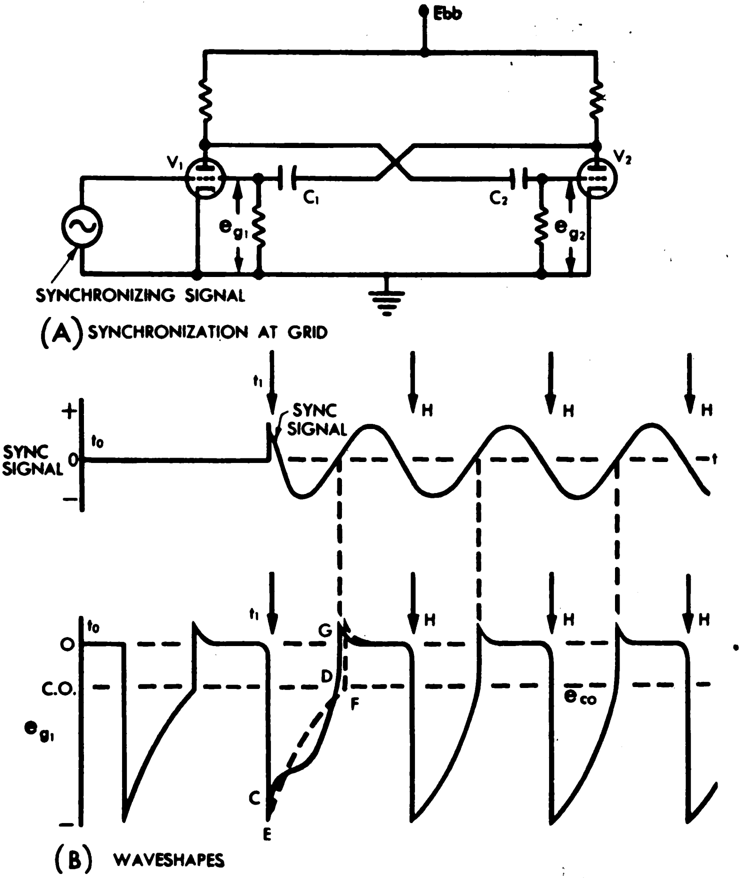


Figure 4–6.—Synchronization with sine wave on grid.

the synchronizing signal drops below zero voltage and starts to decrease the voltage on the grid, the multivibrator regenerative action rapidly cuts off V_1. Since the action occurs at the same time, *H*, on each cycle of the synchronizing signal, the multivibrator is forced to operate at the synchronizing frequency. This latter action occurs in this way only when the synchronizing frequency is higher than the natural frequency of the multivibrator.

The first cycle of the synchronizing signal between t_1 and H will not, in general, be the same shape as the steady-state waveshape. The shape of this cycle depends on the time and phase at which the synchronizing signal is applied. The waveshapes in (B) of figure 4–6 show only one of many possible waveshapes for the first cycle. After one or two cycles, however, the multivibrator adjusts itself to a steady-state condition in which it operates so that the phase relation between the synchronizing signal and the multivibrator output is correct.

SYNCHRONIZATION WITH A POSITIVE PULSE.—Although multivibrators can be synchronized with a sine-wave voltage, a more satisfactory method of synchronization is obtained by the use of short trigger pulses. An application of this circuit is ~~often~~ found in remote or repeater radarscopes. The sweeps of the PPI are generated by a free-running type multivibrator which is locked in by the trigger pulses derived from the radar transmitter. These pulses may be either positive or negative. Note the effect of positive pulses on the multivibrator grid waveform in figure 4–7. A positive pulse, such as at points *B* or *C*, when applied to a tube that is already conducting, serves to cause only a momentary increase in current flow. THUS, A POSITIVE TRIGGER PULSE APPLIED TO A CONDUCTING TUBE OF A MULTIVIBRATOR HAS NO EFFECT ON THE ACTION OF THE MULTIVIBRATOR.

However, when a positive trigger pulse is applied to a nonconducting tube and is of sufficient amplitude to raise the grid above cutoff, as the pulse at *D*, the tubes are switched as current starts to flow in the tube which was formerly cut off. If the trigger pulse occurs at a time such as *A*, the grid voltage will not rise to cutoff as shown at *A′*, and the switching action will not be started. THUS A POSITIVE TRIGGER PULSE APPLIED TO A NONCONDUCTING TUBE IN A MULTIVIBRATOR CAN CAUSE SWITCHING ACTION TO TAKE PLACE ONLY IF THE PULSE IS LARGE ENOUGH TO RAISE THE GRID ABOVE THE CUTOFF VOLTAGE.

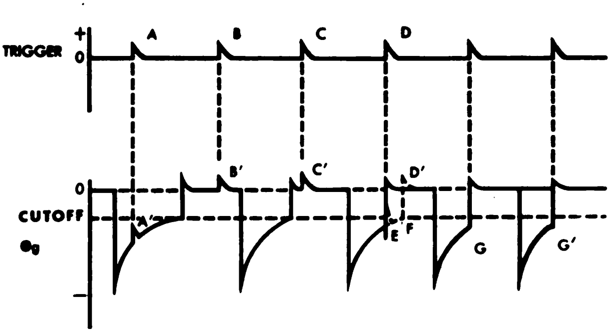


Figure 4–7.—Synchronization with positive pulses on grid.

With the exception of the trigger pulses, the grid waveform illustrated is that of a free-running multivibrator until time *D*. The positive pulses which occur at *D* drive the grid above cutoff, so that the cycle of the multivibrator is shortened by an amount equal to *EF*. In order for proper synchronization to take place, the period of the multivibrator must be greater than the interval between trigger pulses. Then the trigger pulses cause the multivibrator to switch earlier in the cycle than it would if free-running. As a result, the grid has not reached cutoff by time *G* when the next trigger pulse is applied. Thus, the frequency of the multivibrator is forced to be the same as the repetition frequency of the trigger pulses, and the multivibrator will be switched consistently at time *G* after the transition is made from the free-running condition.

SYNCHRONIZATION WITH A SUBMULTIPLE OF TRIGGER FREQUENCY.—A multivibrator may also synchronize to a submultiple of the trigger frequency. As shown in figure 4–8, trigger *A* causes switching of the multivibrator but triggers *B* and *C* have no effect since they are applied to a conducting tube. Trigger *D* is applied to a nonconducting tube, but it is not large enough to start conduction so that the next pulse to cause a switching action is trigger pulse *E*. In the case shown, every fourth trigger pulse switches the multivibrator so that

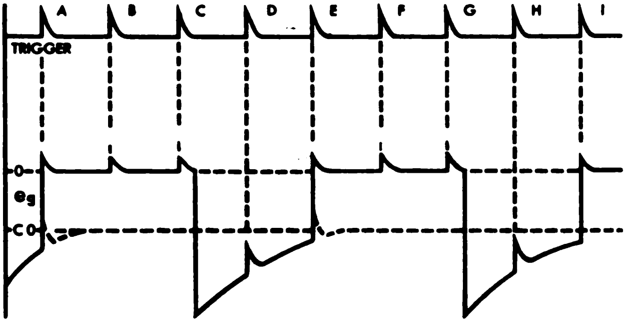


Figure 4–8.—Synchronization at a submultiple of trigger frequency.

the repetition frequency of the multivibrator is one-fourth of the trigger pulses.

By passing the output of this multivibrator through a differentiator a count-down circuit is obtained. This combination of a submultiple synchronized multivibrator and a differentiator produces one pulse for each four trigger pulses. Other multiples of a pulse count-down may be obtained by changing the R–C time constants of the multivibrator.

SYNCHRONIZATION WITH NEGATIVE PULSES.—Negative pulses can be used to trigger a multivibrator as well as positive pulses. As shown in figure 4–9, the first few cycles of the waveforms show the voltage variations that take place at the grids of a free-running multivibrator. As shown at *B* and *B′* and at *C* and *C′*, A NEGATIVE TRIGGER PULSE APPLIED TO A NONCONDUCTING TUBE OF A MULTIVIBRATOR HAS NO EFFECT ON THE OPERATION OF THE MULTIVIBRATOR. However, if the negative trigger pulse is applied to the conducting tube, that tube operates as a single-stage amplifier and applies a larger positive pulse to the grid of the nonconducting tube. The negative trigger pulse *A* reduces the grid voltage of V_1, as at *A′*, causing an increase in e_{g2}, as at *A″*. Since the change of e_{g2} is not sufficient to cause switching, trigger pulse *A* has no effect on the circuit. When a trigger pulse occurs in a later phase relative to the positive alternation of e_{g1}, as at *D′*, the amplified pulse applied to

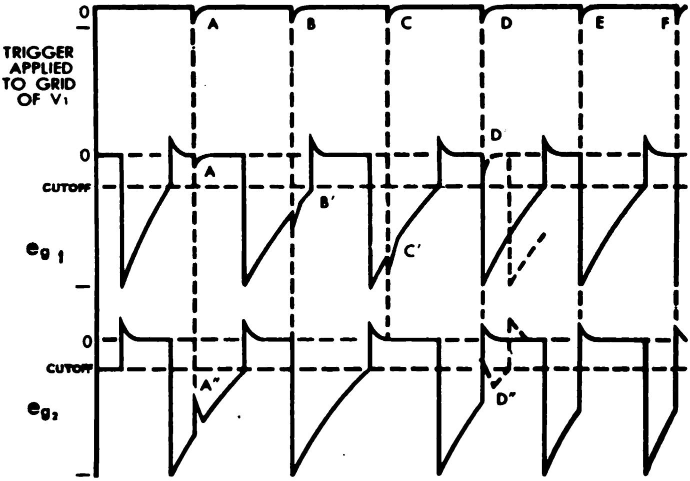


Figure 4–9.—Synchronization at grid with negative pulses.

V_2 at D'' raises the grid above cutoff, exactly as if a positive pulse were applied directly. THUS A NEGATIVE TRIGGER PULSE APPLIED TO THE CONDUCTING TUBE OF A MULTIVIBRATOR CAN SYNCHRONIZE THE MULTIVIBRATOR, PROVIDED THE AMPLIFICATION BY THE CONDUCTING TUBE PRODUCES A RESULTANT POSITIVE PULSE LARGE ENOUGH TO RAISE THE GRID VOLTAGE OF THE NONCONDUCTING TUBE TO CUTOFF. Note that it is not necessary for the negative trigger to reduce the grid voltage of the conducting tube to cutoff.

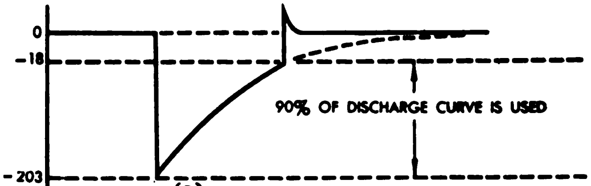
The synchronization voltage can be introduced at the plate as well as at the grid and cathode. The voltage has very little effect on the plate current of the tube to which it is introduced but has the usual effect on the other grid. The coupling capacitor transfers the synchronizing voltage to the other grid. Therefore, introducing a pulse of a given polarity at the plate of one tube has the same effect as introducing the pulse with opposite polarity on the grid of the same tube.

Improving Circuit Action

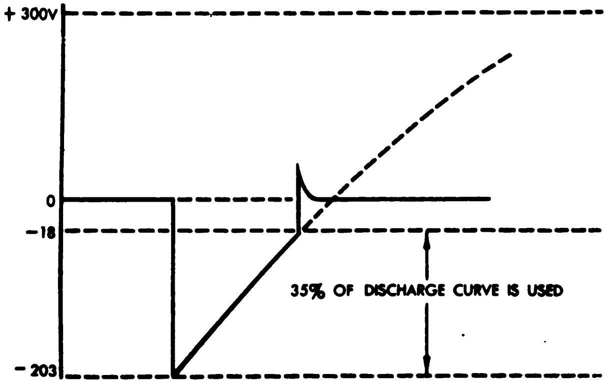
There are a number of slight changes that can be made in the basic plate-coupled multivibrator circuit to improve its stability and output waveshape. One is to return the grids to a positive voltage, and another is to use series-limiting resistors.

RETURNING THE GRIDS TO A POSITIVE VOLTAGE.—Returning the grids to a positive voltage improves the stability of a plate-coupled multivibrator. In this arrangement, the grid-leak resistors are often connected to the source of plate voltage, or the grids are returned to B plus. Under this condition the grid-current flow causes the B-plus potential to exist mostly across the resistor while the grid-to-cathode voltage remains within a fraction of a volt or zero. When the plate voltage of the other tube drops, the capacitor discharges as usual, with a current that is high enough to drive the grid negative by the amount of the plate-voltage drop. Due to this condition the exponential discharge of the capacitor appears as though the capacitor were charged to the plate-voltage drop plus the plate-voltage supply. The net result is that the curve shown in figure 4–10 (A), resulting from the negative peak value to cutoff, is much more linear due to the fact that a smaller percentage of the discharge curve is used.

Returning the grids to a positive voltage produces an advantage in that the discharge curve strikes cutoff at a more nearly vertical angle. This means that if variations of voltage or temperature in the circuit affects the relation of the waveshape to the cutoff value only a small change of time will occur. In contrast, study the grid-voltage curve in (B) of figure 4–10 where the grid resistor is grounded. The discharge curve is almost parallel at the cutoff value. In this case a small variation of the cutoff value will cause a large time change. In conclusion, then, returning the grids to B-plus decreases the change in duration of pulses and the free-running frequency for a given change in circuit constants.



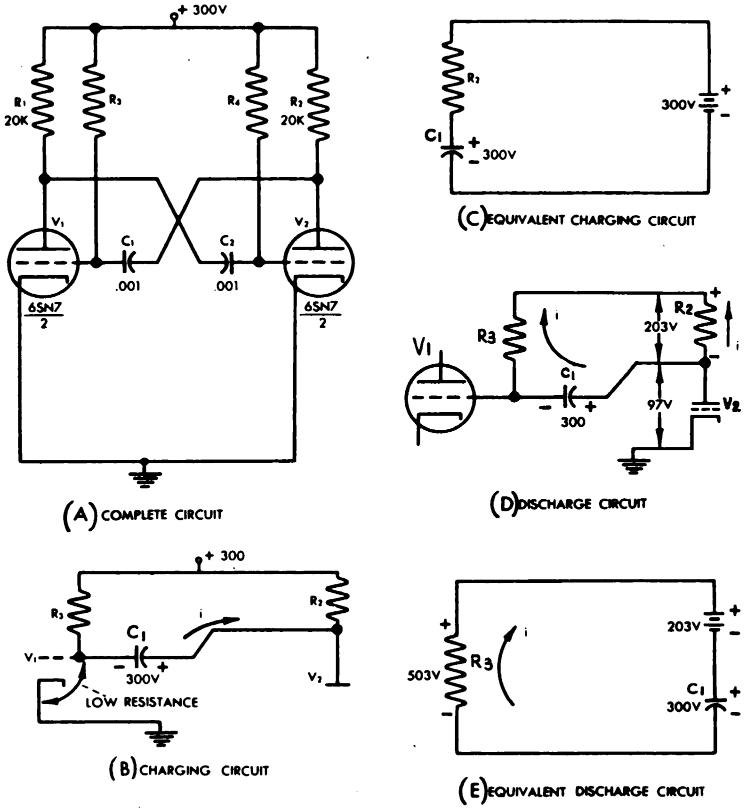
(B) e_{g_1} WITH GRID RETURNED TO GROUND



(A) e_{g_1} WITH GRID RETURNED TO + 300V

Figure 4–10.—Improving the basic circuit.

You can increase your understanding of the effect of returning the grid to B plus by making quantitative analyses and studying equivalent circuits. In figure 4–11 (C), notice the equivalent circuit during the time that V_1 is conducting. In (B), the parts of the circuit involved during the charge of C_1 are shown. Since the grid-cathode resistance is very small, compared to the grid-leak resistor, it can be represented as a direct connection, producing the equivalent circuit in (C). Here R_2 is small and as a result the capacitor will be fully charged before tube V_2 starts to conduct. Thus at the time V_1 is cut off, capacitor C_1 starts with a charge of 300 volts. The discharge circuit for C_1 is shown in (D). V_2 is conducting during this time, causing a voltage of 203 volts across the load resistor R_2. A closed circuit is formed as shown in (E). The voltage across the load resistor may be considered as a source of voltage or battery which is of a polarity to aid the capacitor in



(A) COMPLETE CIRCUIT

(B) CHARGING CIRCUIT

(C) EQUIVALENT CHARGING CIRCUIT

(D) DISCHARGE CIRCUIT

(E) EQUIVALENT DISCHARGE CIRCUIT

Figure 4–11.—Effect of returning grid to B plus.

causing current flow. The total voltage across the resistor R_3 becomes the capacitor charge plus the other voltage (300+203=503 volts). The grid end of resistor R_3 is 503 volts negative with respect to the end connected to 300 volts. As a result, the grid is 203 volts negative with respect to the cathode. Effectively, the capacitor is charged by the plate-voltage source.

Since the polarity of the source is reversed, the plate-voltage source will aid the discharge of the capacitor. When a source is connected to a charged capacitor with a polarity that discharges it, the capacitor will completely discharge and recharge to the new voltage all in one long exponential curve. In this case, the curve will have the same amplitude as it would if it charged from zero to 503 volts. The resulting slope is as shown in figure 4–10 (A). Of course, the slope due to the more rapid discharge will go through zero earlier if the same R–C is used. Therefore a longer R–C must be used when grid resistors are returned to a positive voltage than is used when the resistors are returned to ground, for the same duration of output pulse.

SERIES-LIMITING RESISTORS.—The use of series-limiting resistors in the grid circuits of this multivibrator

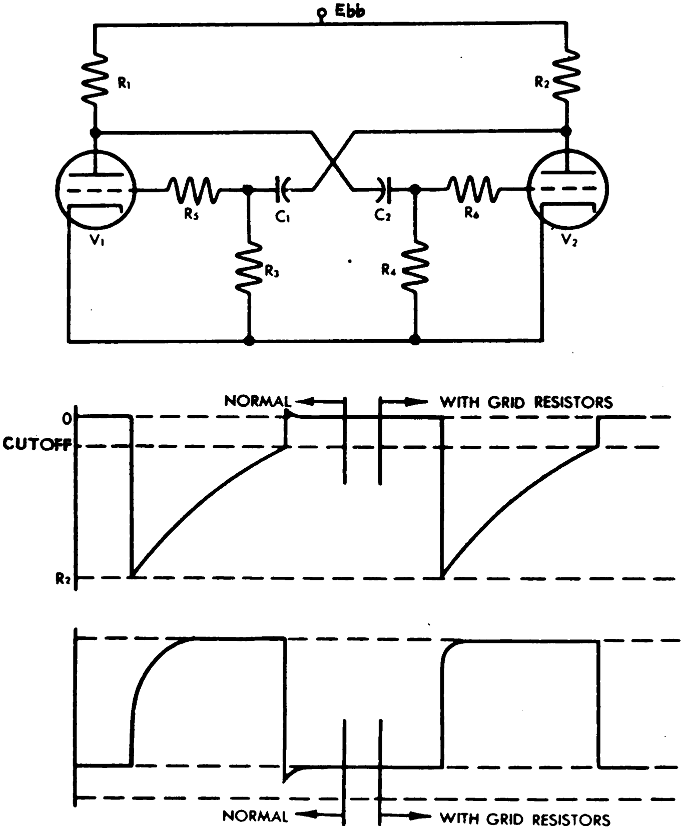


Figure 4–12.—Using series-limiting resistors.

increases the squareness of the waveshapes. In the circuit shown in figure 4–12, these resistors are R_5 and R_6. The first half cycle of the plate and grid waveshapes shown is without the grid-limiting resistors in the circuit, and the second half cycle is with R_5 and R_6 in the circuit. Note that the grid does not go positive when resistors are provided. When the other plate becomes more positive, most of the voltage change occurs across the grid-limiting resistor while only a small fraction of a volt exists between the grid and cathode, provided the grid-limiting resistor is of the order of 500K or more. The plate that is going positive cannot normally rise in voltage any quicker than the coupling capacitor can charge. So in the normal circuit, the plate waveshape resembles the coupling capacitor charging curve. When grid-limiting resistors are inserted, the limiting resistor is in series with the low grid-cathode resistance. These two resistors and the grid-leak resistor form a high-resistance circuit for the capacitor to charge through. The circuit is like the trapezoid sweep generator in that a "jump" voltage will occur across the limiting and grid-leak resistor. Then the exponential rise will start from there. In the waveshape of this illustration, the jump is about 90 percent of the total amplitude, with only slight rounding of the leading edge of the pulse.

The Driven Plate-Coupled Multivibrator

You previously saw how the multivibrator circuit can be used either as a free-running or as a synchronized square-wave generator. The circuit can also be used as a monostable or one-shot multivibrator. This is accomplished by holding the circuit inoperative until a pulse is applied to trigger or drive it into operation for one cycle and then return it to the inoperative condition.

The one-shot multivibrator circuit shown in figure 4–13 is the same circuit that you previously studied except that here 50 volts of bias is placed on the grid of V_1. This bias keeps V_1 normally cut off and permits V_2 to conduct con-

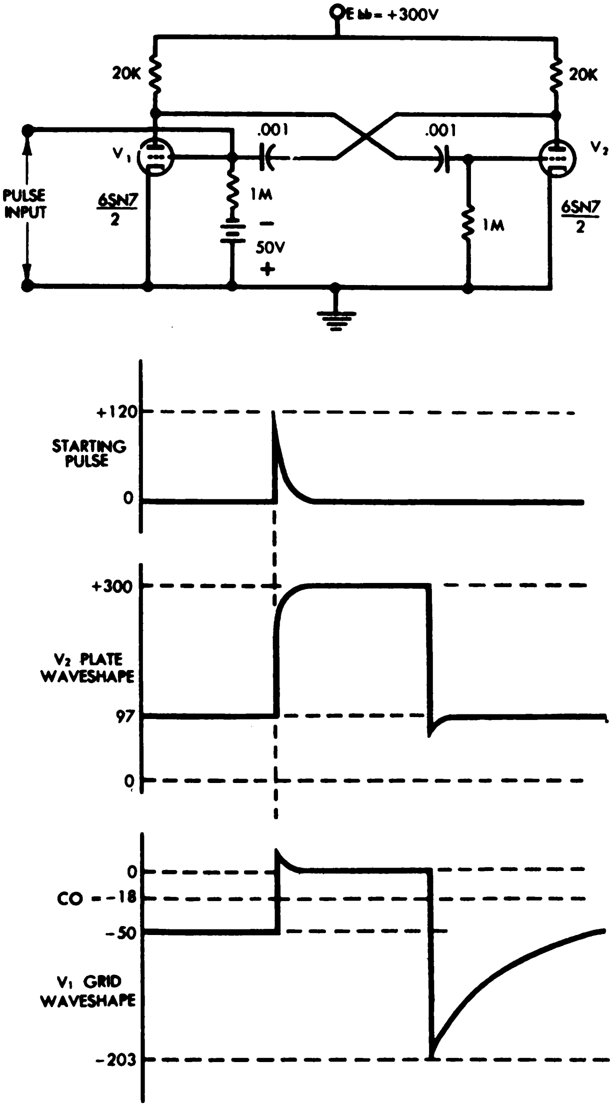


Figure 4–13.—One-shot plate-coupled multivibrator.

tinuously. However, the circuit will operate for one cycle if a voltage is applied to the grid of V_1 with sufficient amplitude to raise the grid voltage above cutoff. Since cutoff is minus 18 volts, the required amplitude must exceed 50 minus 18, or 32 volts positive. Any voltage of more than 32 volts at the grid will cause V_1 to conduct. The resultant drop in plate voltage at V_1 is amplified around the circuit until V_2 is cut off, thus placing the plate of V_2 at supply voltage level. V_2 remains cut off for the usual time and then starts to conduct, driving the grid voltage at V_1 far below cutoff. Because the charge on the V_1 coupling capacitor leaks off to minus 50 volts, which is still below cutoff, the cycle stops as V_1 cannot conduct again. The illustration shows that a positive pulse is applied to the grid to start conduction. One positive square wave is generated at the plate of V_2 for the one pulse. Note in the grid waveshape at V_1 that the V_1 grid voltage starts and ends with minus 50 volts. In order for the circuit to produce another square wave, another pulse must be applied to the grid.

Other methods have been successfully used to keep the driven plate-coupled multivibrator inoperative until pulsed. The circuit in figure 4–14 illustrates one method. This circuit has an advantage over the circuit previously shown in figure 4–13 since a separate power supply is not required. It uses fixed bias provided by a voltage divider from plus 300 volts. With the values shown, the cathode will be 50 volts more positive than the grid. As cutoff is still at minus 18, 32 volts will, as before, be required to make the tube conduct. When conducting starts due to a pulse on the grid, the circuit will oscillate for one cycle, and then quit.

A negative pulse can be used to start a cycle in either of the last two circuits illustrated provided it is introduced at the plate of V_1 or the grid of V_2. This causes the negative pulse to be amplified and inverted by V_2 bringing the grid of V_1 above cutoff and thereby starting tube operation.

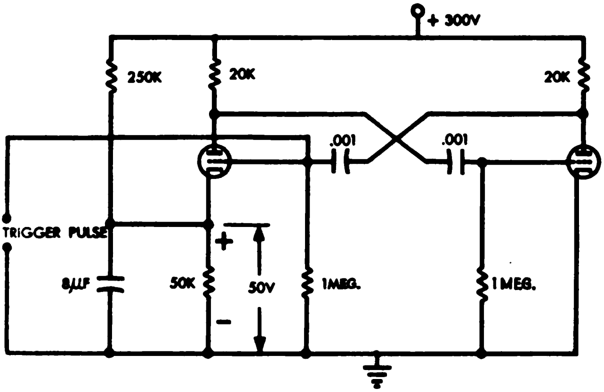


Figure 4–14.—One-shot plate-coupled multivibrator.

Bistable Multivibrator

A modification of the basic plate-coupled multivibrator will give a circuit that has two stable states. Two variations of this circuit are shown in figure 4–15. They are similar to the previously discussed plate-coupled circuit (figs. 4–13 and 4–14), except that a negative bias is applied to the grids of both tubes. This new arrangement requires a separate trigger pulse for each half cycle of operation.

In the flopover circuit, shown in figure 4–15 (A), trigger pulses from different sources are applied to each grid. Once a trigger from one source has caused the circuit to flopover, it then requires a trigger from the second source to generate the other half cycle. In other words, the flopover circuit responds to alternate triggers from the two sources. This type of circuit may be used in the formation of a gate, which has the leading edge formed by a trigger from one source and the trailing edge formed by the trigger from the second circuit.

The scale-of-two circuit shown in figure 4–15 (B) has trigger pulses applied to both grids from a single source. Each pulse drives the circuit from one stable condition to the other. As in the previous flopover circuit, two trigger pulses are needed to generate each complete cycle.

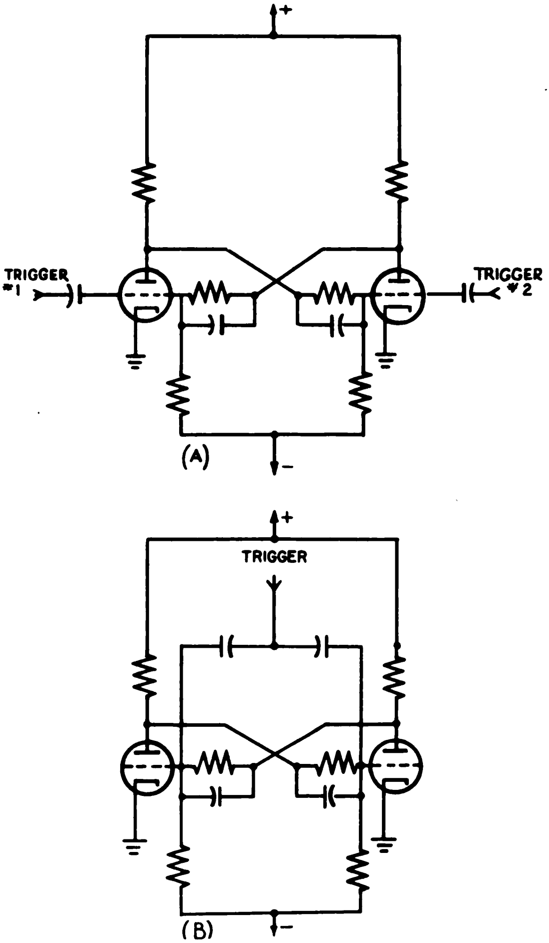


Figure 4–15.—Flopover and scale-of-two multivibrators.

The scale-of-two circuit is sometimes referred to as the BINOMIAL-COUNTER. By using a cascaded group of scale-of-two circuits a scale of four, eight, sixteen, etc., circuits are possible. In other words one pulse will be derived for each four, eight, sixteen, etc., pulses put into the circuit. The scale-of-two circuit is also used for time sharing in video relays. In this case the gate from one tube of the multivibrator is used to turn one video amplifier on and off on alternate pulses and the output from the other multivibrator tube turns a second video amplifier tube on and off at opposite alternations.

The Cathode-Coupled Multivibrator

In the plate-coupled multivibrator, feedback in the correct phase to sustain oscillation was obtained by connecting the plate of each tube to the grid of the other. Feedback in the correct phase can also be obtained by connecting the two triodes as shown in the cathode-coupled multivibrator of figure 4–16. The usual plate-to-grid connection in this case is made from V_1 to V_2, and the cathode of the second tube is capacitively coupled to the cathode of the first. If you follow through these changes, you can see that feedback is in phase. Assume that the current in V_1 is decreasing. As the plate becomes more positive, the V_2 grid will become more positive and current will increase in V_2. This makes the cathode of V_2 more positive. Since the cathode of V_1 is coupled to the V_2 cathode, the cathode of V_1 will become more positive. Because a positive-going cathode is like a negative-going grid, the current in V_1 will decrease. Since the original change was a current decrease in V_1 the feedback is in phase to sustain oscillation.

Notice that in this multivibrator the cutoff time of the grid of the second tube is controlled by the discharge of C_2 through R_2, while the cutoff time for the first tube is controlled by the charge of the cathode coupling capacitor through the cathode resistor of the first tube.

In order to understand the operation of this multivibrator, assume that V_1 is conducting and that V_2 has been cut off by an action similar to that just described. Capacitor C_2 discharges through R_2, R_p of V_1 and R_5, so that the voltage at the grid of V_2 decreases from a high negative potential toward ground as the discharge current through R_2 decreases. When e_{g2} rises to the cutoff voltage of V_2, the tube starts to draw current through R_6. The rising voltage produced across this resistor is coupled through C_1 to the cathode of V_1 since the voltage across the

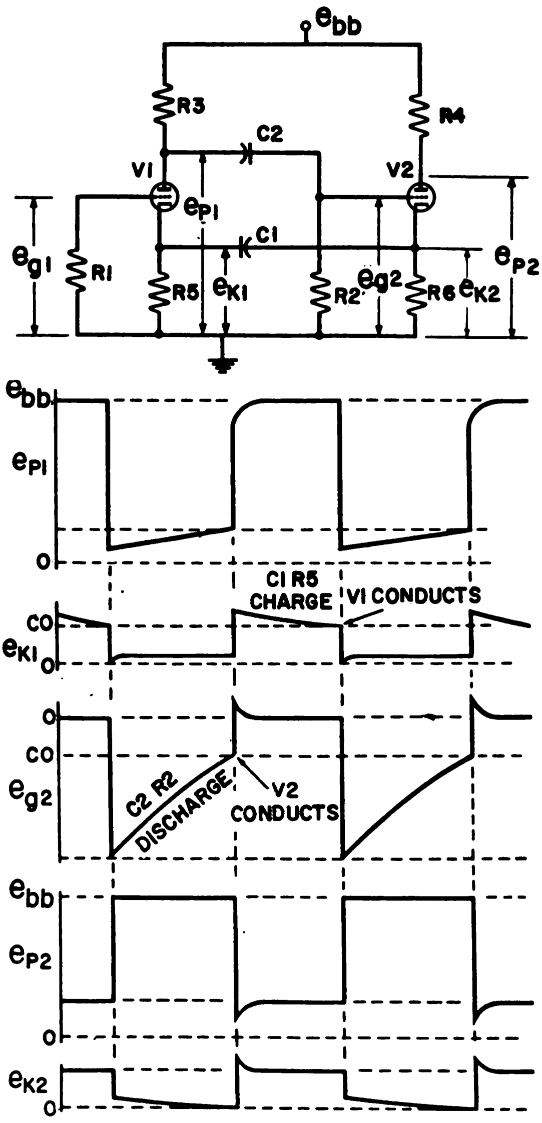


Figure 4–16.—Cathode-coupled multivibrator.

capacitor cannot change instantaneously. The positive-going voltage which is coupled to the cathode of V_1 adds to the existing voltage across R_5, reducing the flow of current in V_1. As i_{p1} decreases, the voltage at the plate of V_1 increases, and this positive-going voltage is coupled to the grid of V_2 to increase further the current in V_2. This action around the circuit in regenerative, and it ends with the current in V_1 reduced to zero and the current in V_2 at a maximum.

Capacitor C_1 is charging to the voltage existing across R_6 through the path R_5,· R_p of V_2, and R_4 to the high-voltage supply. The charging current produces across R_5 a voltage which is more than sufficient to hold V_1 beyond cutoff. However, as the voltage across the capacitor rises, the magnitude of the charging current falls. Consequently, the voltage across R_5 decreases and at some time it will fall below the value required to cut off V_1. At this instant, a current starts to flow through V_1 lowering the voltage at the plate of the tube. This negative-going voltage is coupled through C_2 to the grid of V_2, causing i_{p2} to decrease. As a result of the reduction of i_{p2} and the decrease in e_{k2}, the voltage across R_5 decreases. Therefore, C_1 discharges through R_6, R_p of V_1, and R_3 in an effort to charge in the opposite direction to the voltage across R_5. The cumulative effects of the discharge of C_1 through R_6 and the negative-going voltage applied to the grid of V_2 combine to cut off V_2 almost instantaneously.

The waveshape at the important points are shown in figure 4–16. While V_1 is cut off, a cathode voltage exists due to the charge of the cathode-coupling capacitor. The voltage is visible in the waveshape e_{k1}, as the high initial charging current causes a high cathode voltage. At a certain positive cathode voltage the grid-to-cathode voltage difference will be so great that no plate current will flow. The cathode voltage exceeds this value at first, but as the capacitor charges, the resistor (R_5) voltage decreases exponentially until the cutoff value is reached. Upon dropping below this value, current flows in V_1, which by the drop in e_{p1} cuts off V_2. This discharge of C_2, for holding e_{g2} below cutoff, is the usual action as observed with the previous circuit. During the time V_2 is cut off, its plate voltage rises instantly to E_{bb}, forming a square waveshape. There is no rounding of the corner because there is no capacitor to charge. When V_2 is cut off, the cathode voltage tends to become zero, but it cannot be zero until the capacitor is discharged. Therefore, the waveshape shows first a sudden drop as plate current ceases, then an exponential drop as the capacitor discharges. The same discharge is indicated by the plate and cathode waveshape of V_1. The discharge is such that it makes the cathode of V_1 more negative, so the plate voltage drops below the zero grid-voltage level. As the discharge progresses, capacitor voltage decreases and the effects of it decrease. The plate voltage gradually rises and a slight rise of cathode voltage is seen.

Generally speaking, increasing the size of any part in the circuit except R_1 will lower the frequency. The duration of each half cycle is dependent upon every part in the circuit. Changing any one value will affect the duration of both half cycles. Increasing the power supply voltage will also lower the frequency.

Because of the instability of this circuit when free running, it should be synchronized if it is to operate at a constant frequency or if its action is to be simultaneous with the action of some other circuit. Like the plate-coupled multivibrator, this circuit can be synchronized at either plate, grid, or cathode if the right polarity of synchronization voltage is used. If, for example, you should desire to start conduction in the second tube at a certain time, a positive voltage should be applied to the second grid, first plate, or first cathode at that time. A negative pulse is required at the first grid, second plate, or second cathode to do the same thing. This circuit can also be kept normally inoperative and be made to operate only once for each trigger pulse.

The Cathode-Coupled Monostable Multivibrator

A widely used circuit, which never becomes free running and must always be triggered for each output pulse is the one-cycle or cathode-coupled monostable multivibrator (sometimes referred to as a one-shot or single-kick multivibrator). As shown in figure 4–17, the circuit consists essentially of a two-stage resistance-capacitance-coupled amplifier with one tube cut off and the other conducting normally. The balanced condition of the circuit is established by the arrangement for biasing the tubes. The grid of V_2 is connected to its cathode through the resistor R_2, thus it conducts quite heavily, and the resultant voltage drop across R_k, due to V_2 conducting, biases V_1 to cutoff. When V_2 is not conducting, V_1 cannot be cut off by the self-bias developed across R_k.

The steps in the action of the cathode-coupled monostable multivibrator are as follows:

1. V_1 is cut off initially by the voltage drop produced across R_k by i_{p2}, the plate current of V_2 (time A).
2. V_2 is conducting heavily because its grid is at cathode potential (time A).
3. A positive trigger pulse (time B) sufficient in amplitude to raise the grid of V_1 above cutoff voltage, is impressed on the grid of V_1 through C_1.
4. V_1 begins to conduct and its plate voltage decreases. This decrease passes through C_2, as the voltage across a capacitor cannot be changed instantaneously, and appears on the grid of V_2 as a negative-going voltage.
5. The negative-going voltage on the grid of V_2 decreases i_{p2}.
6. The voltage drop across R_k decreases, allowing more current to flow in V_1.
7. The plate voltage of V_1 is still further decreased.
8. The grid of V_2 goes more negative.
9. This action is repeated until V_2 is cut off and then V_1 is conducting heavily. The switch from V_1 to V_2 is practically instantaneous.
10. V_2 begins to conduct (time C).

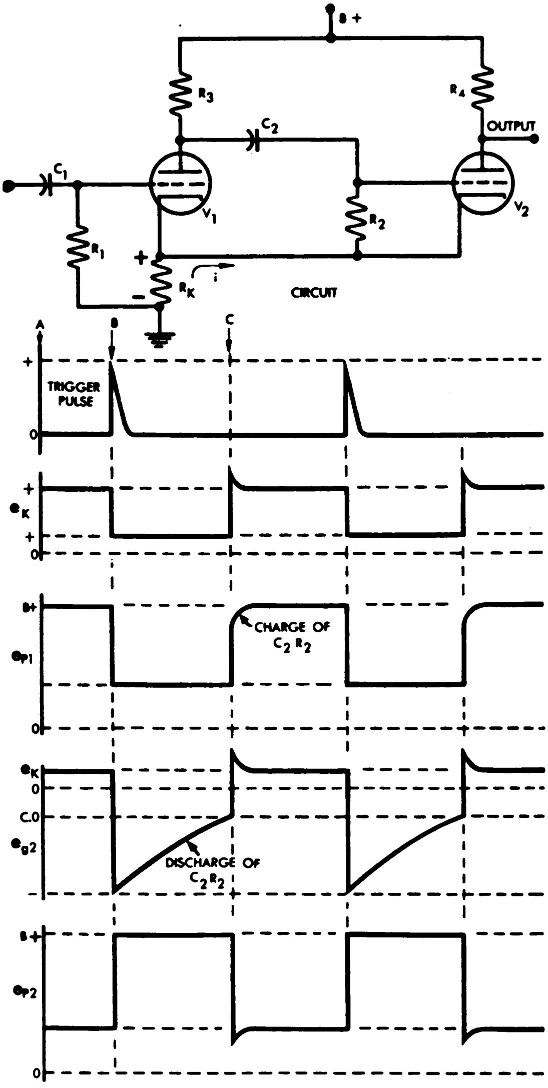


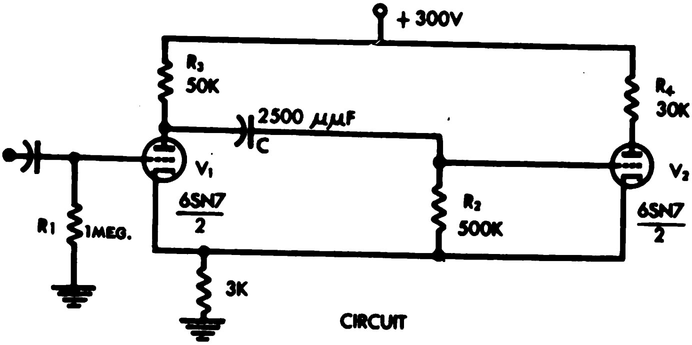
Figure 4–17.—Cathode-coupled monostable multivibrator.

11. The plate current of V_2 flowing through R_k, raises the cathode voltage of V_1, thus reducing its plate current.
12. The decreased plate current of V_1 allows the plate voltage of V_1 to increase.
13. This increase is coupled to the grid of V_2, increasing still further its plate current.
14. The action described is repeated until V_1 is cut off and V_2 is conducting heavily. This action also is practically instantaneous.

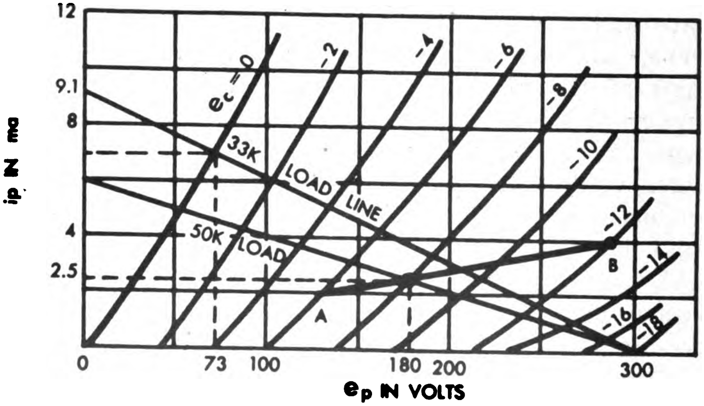
The circuit has now come back to its original balanced state and will remain so until another positive pulse arrives and causes V_1 to conduct. When applied to the input of the monostable mulitvibrator, every positive trigger pulse which causes V_1 to conduct, results in a large positive-pulse output from the plate circuit of the second tube. The length of the positive-output pulse produced at the plate of V_2 is controlled by the time constant of C_2 times R_2. If larger values of C_2 and R_2 are used, the length of the positive output is increased. A positive-output voltage pulse is produced for each positive-input trigger pulse.

QUANTITATIVE ANALYSIS.—The quantitative determination of the amplitude and duration of the output pulse is not too difficult to accomplish. (See fig. 4–18.) The circuit shown is the most basic one. The waveshapes are similar to those in figure 4–17 except that the voltages and time intervals have been determined and the waveshapes drawn to scale.

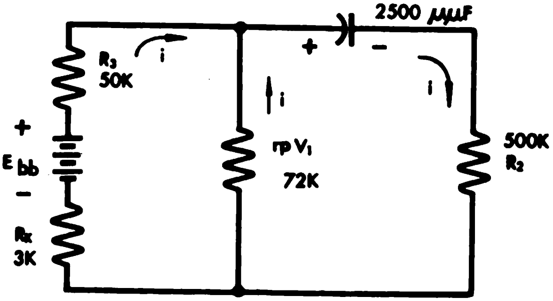
Start by finding the voltages in the circuit before the trigger pulse arrives. Check to see that V_1 is cut off with the 3-K resistor. The circuit will not work if R_k is too small. V_2 is conducting due to zero bias. Now draw a loadline on the characteristic curve to find the voltages. In drawing the loadline, consider the cathode resistance. Any resistance in series with the tube will affect the current, but if the cathode resistor is quite small, its effect is negligible. You can obtain reasonably accurate results



CIRCUIT



6SN7 CHARACTERISTICS



EQUIVALENT DISCHARGE CIRCUIT

Figure 4–18.—Quantitative analysis of cathode-coupled monostable multivibrator.

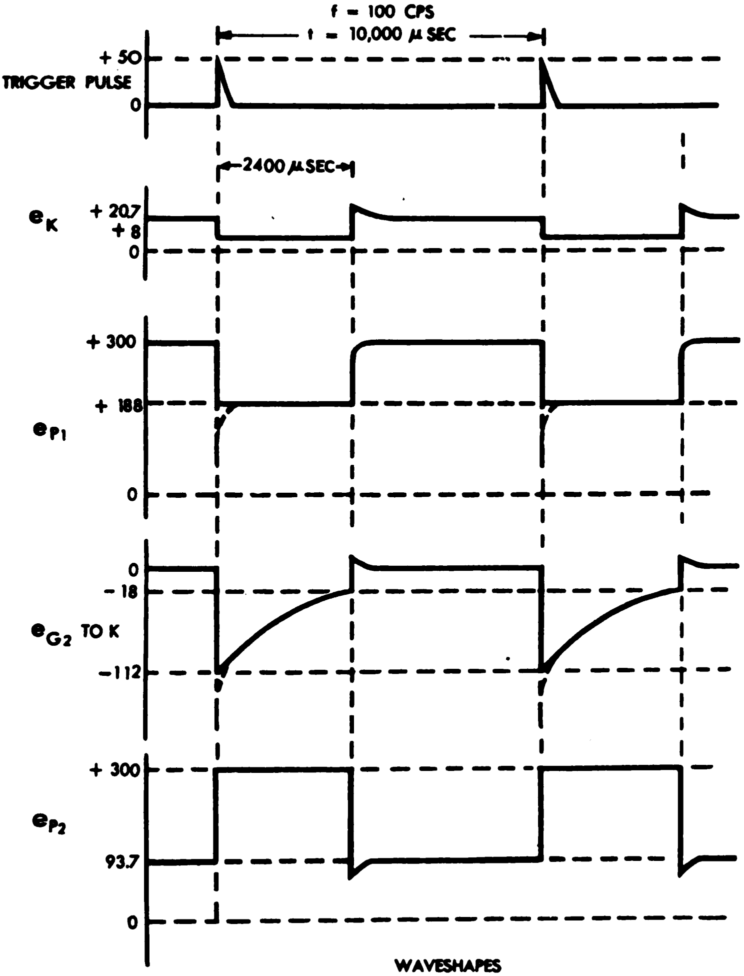


Figure 4-18.—Quantitative analysis of cathode-coupled monostable multivibrator. Continued.

by disregarding the cathode resistance when it is less than 10 percent of the load resistance. When it is 10 percent or more, sketch the loadline for the total cathode and load resistance. In the case of V_2 in this problem, the cathode resistance is exactly 10 percent of the load resistance; so the loadline is drawn for 30K+3K=33K. Check the plate current and plate voltage at the intersection of the loadline and zero grid bias. The plate voltage appears to be 73 volts and the current 6.9 ma. Knowing the current through R_k, find the voltage across it. ($E=IR$, $i_pR_k=0.0069\times 3{,}000=20.7$ volts.) Cutoff is 18 volts, as read from the curves, so the first tube is actually cut off, and the circuit will function.

Now proceed with the calculations. The waveshape shown is the plate-to-ground voltage while the characteristic curves show plate-to-cathode voltage. So you must add the cathode voltage above ground to the plate voltage to find the actual plate-to-ground voltage. Since the cathode is 20.7 volts above ground, add this to 73 to get 93.7 volts at the plate of V_2. To simplify the picture, the waveshape at the grid of V_2 shows the voltage at the grid with respect to the CATHODE rather than ground. In the waveshape of cathode voltage shown, notice that before the trigger pulse arrives, the cathode voltage is 20.7 volts as previously calculated.

Now, determine the voltages in the circuit immediately after the trigger pulse and switchover has occurred. During this time, the second tube is cut off so the plate voltage there is 300 volts. The first tube is conducting as much as a self-bias 3-K resistor will allow. To find this value, a loadline and bias line are required. Since 3K is less than 10 percent of 50K, the loadline can be drawn for 50K only. For the bias line, assume different values of current, find the bias, and plot the points on the characteristic curves. For example, if the current is 2 ma., the bias would be i_pR_k, or 0.002 × 3,000, or 6 volts. To plot this, find the 2 ma. current line and the 6-volt grid-voltage curve (point *A*). Assume a second current of 4 ma. The bias is 12 volts if 4 ma. flows through the 3-K resistor. Mark the intersection of the 4 ma. current line and 12 volts grid-voltage line (point *B*). Draw a line through these points and where the bias line intersects the loadline is the operating point for the tube. According to the illustration, it appears that the bias or cathode voltage is 8 volts. The plate-to-cathode voltage is 180 volts. So the plate voltage has changed from 300 to 180 plus 8, or 188 volts. The change is 112 volts in the negative direction. The capacitor must discharge by this amount through the 500-K grid resistor R_2. The new plate voltage at V_1 is shown on the E_{p1} curve.

The grid voltage at V_2 goes from zero to negative 112 volts at the switchover time, then decreases exponentially toward zero. This holds V_2 nonconducting until the voltage decreases below cutoff. From the curves, cutoff is minus 18 volts. In order to calculate the time of this exponential decrease, the *RC* must be known. *C* is given as 2,500 $\mu\mu$f. The grid resistor of 500K can be considered as the total *R* for approximate calculations, but the current flows through the 500-K resistor, then through two parallel paths in returning to the capacitor. One path is through R_k, the power supply and R_3 in series. The other path is through the tube. Assuming the internal resistance of the power supply is zero, the first path has a resistance of 53K. The other parallel path, through the tube, has a resistance which is somewhat variable due to bias change during the discharge, but is quite near the value determined by dividing the plate voltage by the plate current. From the curves, the plate voltage is 180 volts, the plate current 2.5 ma., both at the operating point with 8 volts bias. Since $r_p = e_p/i_{p1}$, the tube resistance is 180/0.0025, or 72K. This parallel combination (with a net resistance of 30.5K) is in series with the 500-K resistor to form an equivalent circuit as shown in the equivalent discharge circuit. From this equivalent circuit you can see that the values for determining time constant are as follows:

$$e_{g2} \text{ cutoff} = -18 \text{ volts}$$

$$e_{g2} \text{ max neg} = -112 \text{ volts}$$

$$R = 530.5K$$

$$C = 2{,}500 \ \mu\mu\text{f.}$$

The capacitor will discharge from minus 112 volts to minus 18 volts, which represents a drop of 112 minus 18 or 94 volts. The percentage of discharge equals 94/112 or 84 percent, thus the capacitor will have 100 minus 84 or 16 percent of full charge left when V_2 starts to conduct. From the universal time constant chart, for 16 percent of full charge it is found that 1.82 *RC*'s are required. One $RC = 530.5K \times 2{,}500\ \mu\mu f.$ or 1,320 microseconds. Then $1.82\ RC\text{'s} = 1.82 \times 1{,}320 = 2{,}400\ \mu sec.$ The positive square wave at the second plate will have a duration of 2,400 microseconds and its amplitude will be 300 minus 93.7, or 206.3 volts.

The circuit switches over again at the end of 2,400 microseconds. V_2 conducts, raising the cathode voltage to 20.7 volts, cutting off V_1. The plate voltage of V_1 goes to 300 volts as fast as the capacitor can charge. The capacitor charging current raises the grid of V_2 to a positive voltage momentarily and the cathode above 20.7 volts momentarily. This in turn causes a momentary dip in the plate voltage at V_2. The amplitude and duration of these momentary charges can be calculated by using the same procedure outlined for the plate-coupled multivibrator.

A trigger pulse of a large voltage will cause the dip shown by dotted lines in the waveshape of V_1 plate and V_2 grid in figure 4–18. This dip is an amplified version of the trigger pulse and lasts no longer than the trigger pulse. It does not affect the discharge time of the capacitor. If the pulse is of the minimum amplitude required to trigger this circuit, this dip will hardly be noticeable. Experience has shown that the trigger pulse should be of the lowest possible amplitude consistant with reliable operation.

The duration of the output pulse can be varied from time duration of the trigger pulse to the time between pulses. If attempts are made to make the pulse longer than the time between pulses, the second pulse may not start a new cycle, and the circuit would operate on every other trigger pulse.

EFFECT OF VARYING CIRCUIT ELEMENTS.—The usual method of varying the output pulse width is to vary the size of the *RC* in the grid of V_2. The pulse width will increase with the resistance of R_2 and the capacity of C. Another result is that the amplitude of the output pulse will increase with an increase of the plate-load resistor. Notice these effects in the chart in figure 4–19 which summarizes the effects of varying the circuit elements in the monostable multivibrator.

The stability of the pulse width from this circuit is improved by returning the grid of V_2 to B plus. The capacitor discharges at a higher and more nearly linear rate below the cutoff value, causing the slope to be steeper at cutoff.

Although the trigger pulse is applied to the first grid in this circuit, the circuit can also be triggered at the plate of the first tube or the grid of the second tube if a negative pulse is used. The action will be the same.

The monostable multivibrator (fig. 4–20) is used in a radar set to produce a pulse of measured duration which drives a sawtooth sweep generator. The basic circuit is modified several ways to produce this more efficient circuit. A third tube is incorporated in the circuit to isolate the multivibrator from the source of trigger voltage and to amplify and sharpen the pulse. This third tube (V_3) is biased beyond cutoff so that it acts as an open circuit at all times except when a trigger pulse is on the grid. The grid of the first tube (V_1) has about plus 14 volts on it from the plus 300-volt voltage divider, while the grid of the second tube (V_2) is returned to plus 300 volts through a large resistor. The current through V_2 and the cathode resistor is high enough .to cut off the first tube in spite of the positive voltage on V_1 grid. This positive voltage partly determines the duration of the output pulse. The circuit operates in this way—a positive trigger pulse, simultaneous with the transmitter pulse, is applied to the grid of V_3. This tube is called a CLIPPER since a 22-volt bias keeps it cut

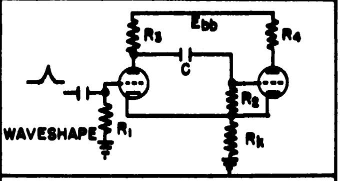
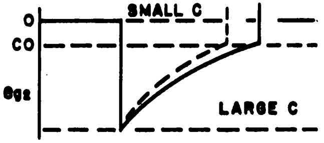
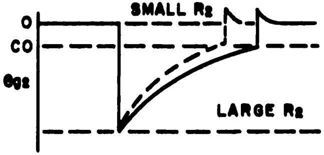
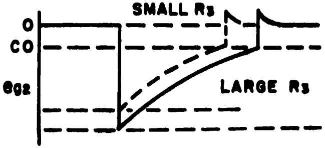
CIRCUIT CHANGE	EFFECT ON DURATION OF OUTPUT PULSE	EFFECT ON AMPLITUDE OF OUTPUT PULSE	EFFECT ON FREQUENCY OF PULSE REPETITION	EFFECT ON STABILITY	
INCREASE C	INCREASED BECAUSE DISCHARGE LASTS LONGER	NO CHANGE	NO CHANGE FREQUENCY GOVERNED BY TRIGGER FREQUENCY	NO CHANGE SAME PERCENTAGE OF RC USED	
INCREASE R_2	INCREASED DISCHARGE IS SLOWER	NO CHANGE	NO CHANGE	NO CHANGE SAME PERCENTAGE OF RC USED	
INCREASE R_3	INCREASE BECAUSE AMPLITUDE OF e_{g2} IS DECREASED	NO CHANGE	NO CHANGE	DECREASED BECAUSE DISCHARGE CURVE HAS LESS SLOPE AT CUTOFF	

Figure 4–19.—Varying elements in monostable multivibrator.

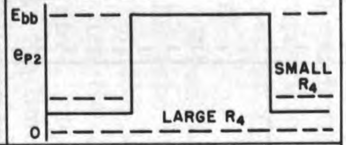
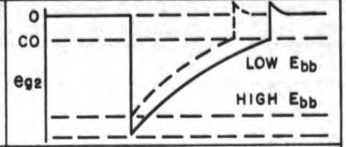
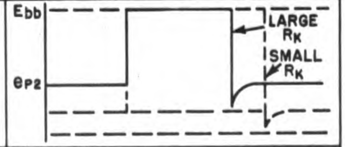
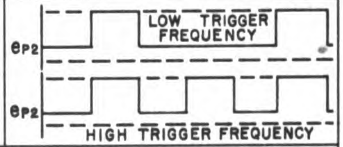
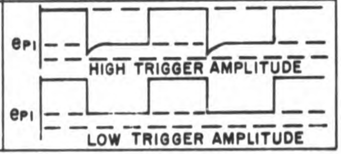
INCREASE R_4	NO CHANGE DOES NOT APPRECIABLY EFFECT PRECEDING CIRCUIT	INCREASED SINCE C_{g2} IS LOWER BETWEEN TRIGGER PULSES	NO CHANGE	NO CHANGE	
INCREASE E_{bb}	INCREASED BECAUSE V_1 PLATE SWING AND V_2 GRID VOLTAGE CHANGE IS GREATER	INCREASED DUE TO GREATER V_2 PLATE VOLTAGE CHANGE	NO CHANGE	DECREASED SLIGHTLY, LARGER PERCENTAGE OF DISCHARGE CURVED USED	
INCREASE R_k	SHORTER SINCE VOLTAGE CHANGE AT PLATE OF V_1 WILL BE DECREASED	DECREASED DUE TO LESS V_2 PLATE VOLTAGE	NO CHANGE	DECREASED BECAUSE VOLTAGE CHANGE AT PLATE OF V_2 WILL BE DECREASED	
INCREASE TRIGGER FREQUENCY	NO CHANGE UP TO POINT WHERE DURATION IS TIME BETWEEN PULSES	NO CHANGE	INCREASED	NO CHANGE	 eP2 LOW TRIGGER FREQUENCY eP2 HIGH TRIGGER FREQUENCY
INCREASE TRIGGER AMPLITUDE	NO CHANGE	NO CHANGE	NO CHANGE	SMALL INCREASE WILL IMPROVE STABILITY IF ORIGINALLY TOO LOW	 eP1 HIGH TRIGGER AMPLITUDE eP1 LOW TRIGGER AMPLITUDE

Figure 4–19.—Varying elements in monostable multivibrator. Continued.

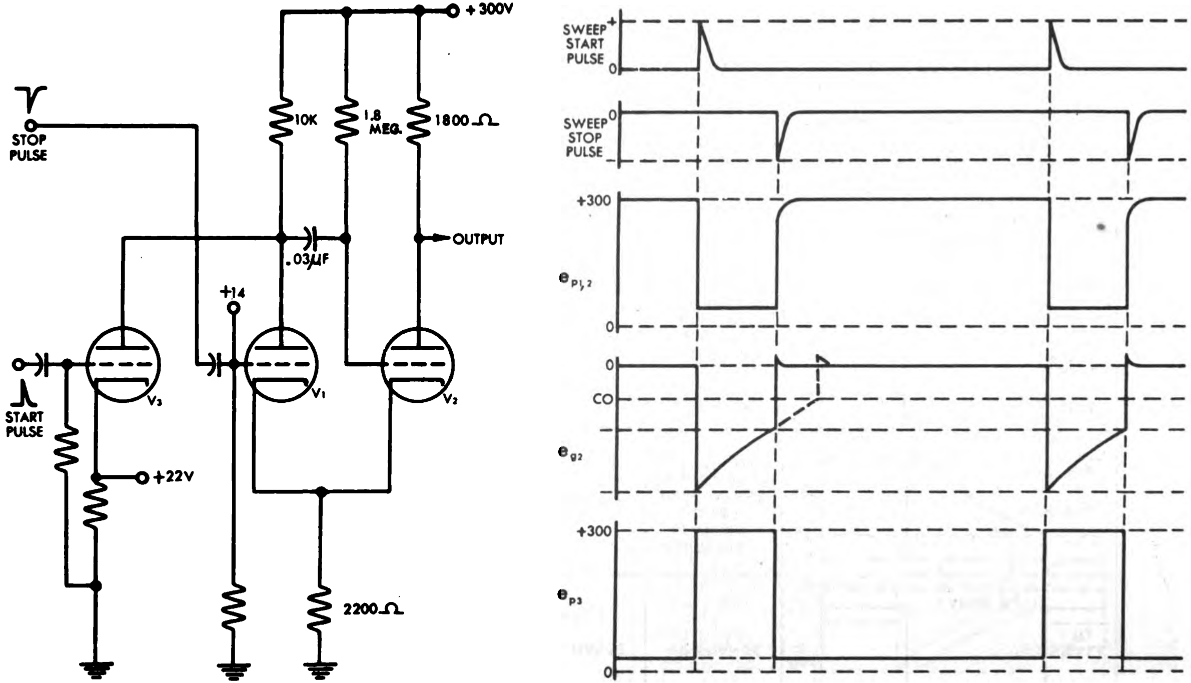


Figure 4–20.—Application of monostable multivibrator.

off and only the peak of the trigger pulse causes conduction. Due to amplification in V_3, a large negative-going charge appears in the plate circuit. This is coupled through the 0.03-μf. capacitor to the grid of V_2. The current in V_2 is reduced, V_1 conducts, the added drop cuts off V_2, and the discharging capacitor keeps V_2 cut off.

When the capacitor discharges down to cutoff for V_2 this tube would conduct, ending the output pulse. However, this is the sweep circuit and the pulse should end when the spot reaches the range mark for maximum range. So a voltage is fed back to this multivibrator at the time the spot reaches the last range mark. This voltage is a negative pulse, applied to the grid of V_1. This reduces the current in V_1, and the small change is amplified around the circuit to cutoff V_1 and causes V_2 to conduct again. Dotted lines in the V_2 grid waveshape show the normal changeover time, which is always later than the sweep-stop pulse. This system of stopping the sweep multivibrator with a voltage from the sweep deflection circuits simplifies the adjustments in the sweep circuit. When the range switch is positioned to change the duration of the sweep, no part of this multivibrator is changed. But the speed of the spot movement is changed so it reaches the last range mark at a different time and the stop-pulse cuts off the multivibrator at a different time. Thus, the duration of the pulse is determined by the sweep circuit rather than the constants of the multivibrator circuit.

You recall that a hard-tube sawtooth generator must be driven with a negative square wave at its grid. The positive square-wave output from this multivibrator is amplified and inverted by a triode amplifier, then is applied to the sweep-generator tube in the normal way.

USE AS A DELAY CIRCUIT.—In addition to producing pulses of a certain duration, the monostable multivibrator is sometimes used to "delay" a pulse. For example, in one radar set, the process of causing the transmitter to

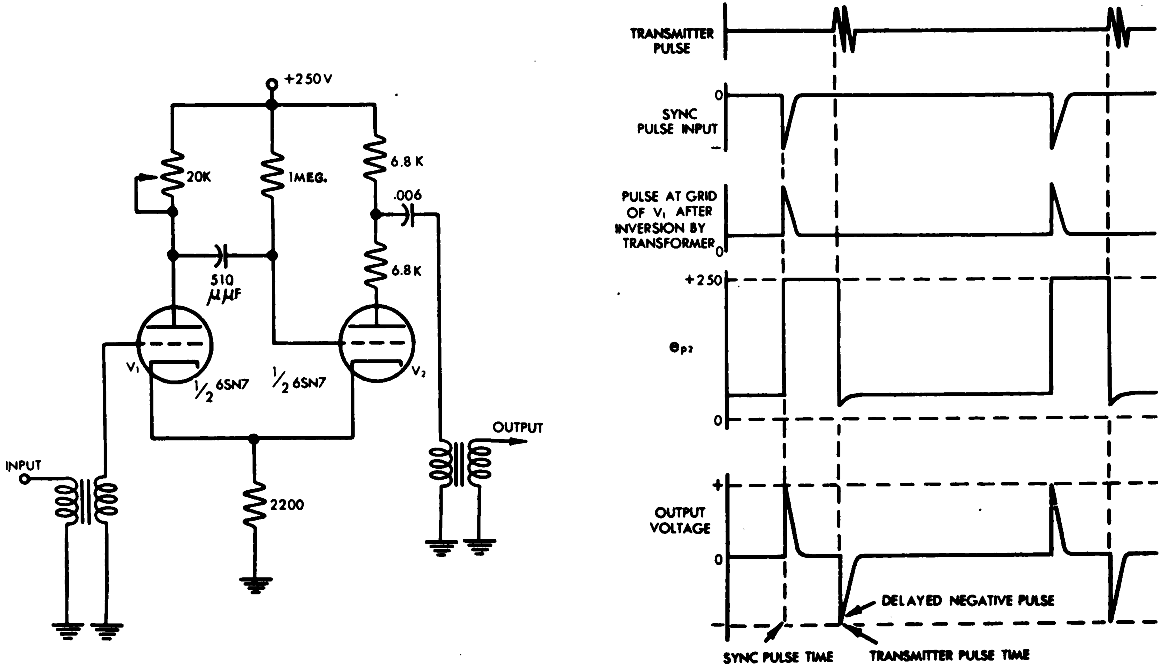


Figure 4–21.—Monostable multivibrator as delay circuit.

operate consumes about 5 microseconds, and the transmitter pulse occurs 5 microseconds after the original trigger pulse. While the sweep voltage in the indicator circuit must start at the same time as the transmitter pulse, the original pulse is used for the sweep voltage so that both the sweep and transmitter pulses occur at the same frequency. However, since the original synchronization pulse would start the time base 5 microseconds before the transmitter operates, it cannot be used directly. Instead, it must be "delayed" 5 microseconds. The monostable multivibrator can introduce the delay. The usual multivibrator circuit is used, with a transformer input, as shown in figure 4–21. The input pulse is negative and occurs five microseconds before the transmitter signal. Due to the transformer connections, the polarity is reversed and the circuit is triggered by the resulting positive pulse. The usual square wave appears at the second plate, with the duration accurately measured by the R–C circuit at the grid. The transformer features a short L/R time constant, so the square wave is differentiated in the secondary. Again by selecting proper secondary connections, the second peak can be made negative.

So the original negative peak is replaced by a similar negative peaked pulse which is delayed 5 microseconds. The delay need only be adjustable over a narrow range, so the plate-load resistor of the first tube is variable.

The Phantastron Circuit

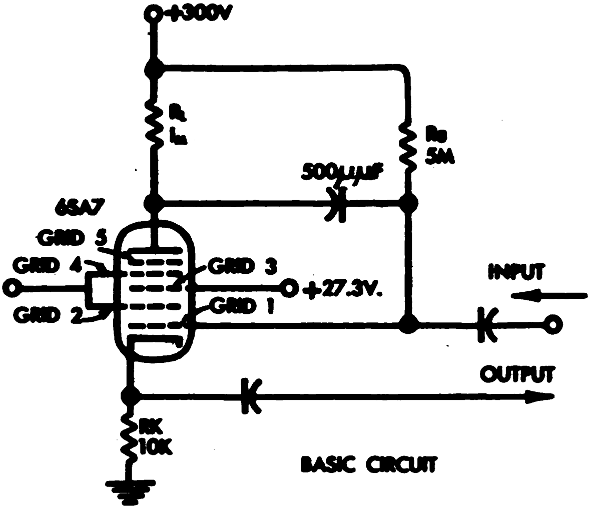
The phantastron is another circuit which is used to delay a timing pulse. It is classed as a medium precision delay circuit. Its operation and output are very similar to those of the monostable circuit. Under power supply voltage changes it is quite stable because its operation depends on the fact that there are specific d-c voltage relationships between the tube elements. Any variation of source voltage varies all these voltages in the same proportion, causing a minimum change in the voltage relationship since all voltages are supplied by voltage dividers across the same voltage source. The phantastron has the unique advantage over multivibrator circuits in that the pulse width, or delay, varies directly with one of the d-c applied voltages. This control by a d-c voltage makes remote control feasible at any distance through unshielded power cable since there is no signal present in the control circuit.

Study the basic phantastron circuit in figure 4–22. This circuit is never free running and is usually triggered by a negative pulse at the control grid (grid No. 1). Consider first the conditions before the trigger pulse arrives. Grids 2 and 4, which are tied together inside the tube, are connected to a voltage divider from plus 300 volts. The normal output of the divider is 86 volts. But with zero control grid voltage and 86 volts on grid 2, 4 ma. of current will flow to grid 2. This additional current in the voltage divider drops grid 2 to 68.7 volts. Grid 3 is connected to a voltage divider which sets its voltage at 27.3 volts. This grid does not draw current because it is negative with respect to the cathode. The current through the 10K cathode resistor is 4 ma. Since the cathode is 40 volts above ground, the grid 3 to cathode voltage is 40 − 27.3 = 12.7 volts. When grid 3 is more than 12 volts negative with respect to the cathode, it prevents electrons from passing it in their journey to the plate. Therefore, since the minus 12.7 volts is beyond plate-current cutoff for grid 3, no electrons will reach the plate. Thus, the plate voltage is plus 300 volts. The grid is connected to E_{bb} through a 5-megohm resistor. A very small grid current flows through the cathode resistor, but the grid voltage may be considered the same as the cathode voltage. In other words, the difference in potential between grid 1 and the cathode voltage is zero. Grid 5, the suppressor grid, is connected to the cathode and has no part in the operation of the circuit other than performing the normal function of a suppressor grid.

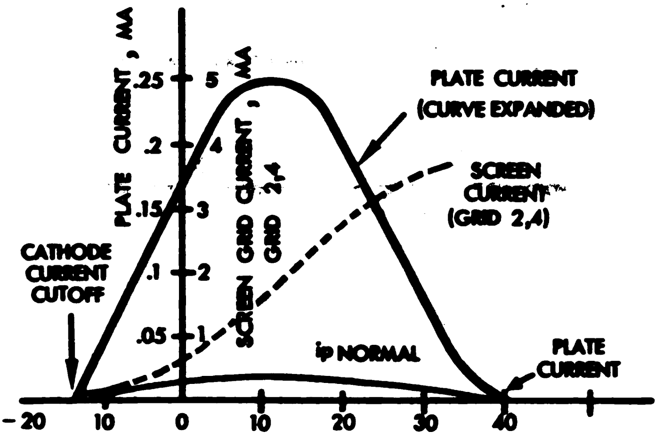
The next factor to consider is the similarity between the phantastron and the start-stop circuit. You can consider the 6SA7 phantastron tube as a tube within a tube. The cathode, grid 3, and plate are equivalent to the first triode in the start-stop circuit. No plate current flows to the plate and the "first tube" is cut off. The cathode, grid 1, and grid 2 form another triode which is comparable to the second tube of the flip-flop multivibrator. As mentioned previously, this triode is conducting. Consequently, any voltage change, which will cause plate current to flow in the triode that is cut off, will start a regenerative switching action.

You can best understand the action of the phantastron when it is triggered by dividing it into three successive phases. The first phase is an extremely rapid change which ends at a first balance point. The second phase is a slow linear change which ends at a second balance point. The third phase is the recovery of the circuit to the pretrigger condition.

The trigger pulse must have the proper polarity and amplitude to bring grid 3 above cutoff. Since grid 3 is 0.7 volt beyond cutoff, 1 volt would do it. A negative pulse of 10 volts amplitude is shown. When this is applied to grid 1, the grid 1 voltage will drop to 30 volts. Cathode follower action will cause the cathode voltage to drop to 30 volts along with the grid. At this point, examine the relation between the cathode and grid 3. Their voltage difference is 27.3 minus 30, or minus 2.7 volts. Grid 3 is not negative enough to prevent electron flow to the plate and plate current will flow as soon as this grid-to-cathode voltage drops below 12 volts. Regenerative action starts with this plate-current flow. The plate current through R_L causes a drop of plate voltage. Grid 1 is capacitively coupled to the plate, so the grid becomes more negative. This drops the cathode voltage still more, which in turn further reduces the voltage difference between grid 3 and cathode. The current to the plate increases, the plate voltage drops, and the grid is driven farther in the negative direction.



BASIC CIRCUIT



6SA7 CHARACTERISTIC FOR ABOVE CIRCUIT

Figure 4–22.—Phantastron circuit and waveshapes.

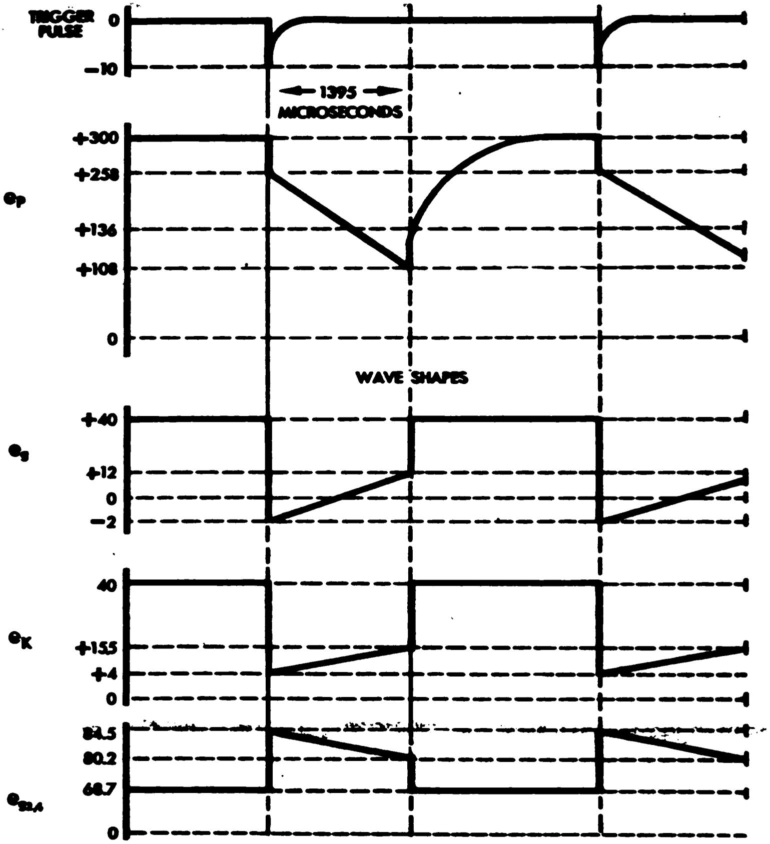


Figure 4–22.—Phantastron circuit and waveshapes. Continued.

Note that the plate current increase is NOT an overall increase in current with a negative-going grid. The negative-going grid decreases the total current slightly. But the resultant decrease in cathode voltage lowers the voltage between grid 3 and the cathode, which allows current to flow to the plate at the expense of the screen current. Grid 3 controls the DIVISION of current between the plate and screen grid. Grid 1 controls the total current. So grid 1 decreases the total current slightly, while grid 3 allows the plate current to increase and causes the screen current to decrease quite rapidly. The 6SA7 characteristic curve (fig. 4–22) shows the relation. In plotting these curves a d-c voltage was applied to grid 1 of the circuit in the illustration and varied from minus 20 volts to plus 40 volts. The dotted curve shows the resultant screen current while the other curve shows the plate current. Note that plate current increases by a small amount while screen current decreases a large amount as G_1 goes negative. The sum of the two currents is the cathode current. It is decreasing because the small current increase is more than overcome by the large decrease.

The regenerative effect just explained is similar to the cutting off of the first multivibrator tube by the second tube. The cutting off process, however, does not go that far in the phantastron, for if the current to the plate were stopped, the plate voltage would have to rise, and this action would be the opposite to the desired feedback. The increase of current to the plate ends at the first balance point. This balance occurs when the negative-going grid 1 voltage decreases the overall current so much that the plate current cannot continue to increase. Grid 3 is unable to divert enough electrons to the plate to maintain the increase of plate current, so the plate current stops increasing and becomes steady. This occurs when grid 1 has dropped to a minus 2 volts, and the plate has dropped 42 volts before the balance point is reached. Each of these voltages is shown in the curves, under the trigger pulse. The grid voltage has dropped to minus 2 volts. Since this is a 42-volt change, caused by the plate, the plate curve shows a 42-volt drop to 258 volts. The cathode voltage drops, due to the decrease in screen current, until with minus 2 volts on the grid, the current is 0.4 ma. The voltage at grids 2 and 4 has gone up because the screen current is decreased, reducing the drop across the voltage divider and raising its voltage to 84.5 volts.

Now consider the action resulting from the starting of the second phase. The capacitor from plate to grid at this time starts to discharge to the new lower plate voltage. At this time the grid voltage is minus 2 and the capacitor discharges from 300 to 260. In discharging, the grid end of the grid resistor becomes more positive. This increases the plate current, starting another drop in plate voltage. Grid 3 is not involved directly in this stage as it is straight triode action between the plate, grid 1, and the cathode that affects the circuit operation. The plate-voltage drop has an effect opposite to that of the discharging capacitor and thereby is degenerative in action. The capacitor discharges, the grid goes POSITIVE, plate current increases, plate voltage drops, tending to make the grid go NEGATIVE. But the plate cannot drop enough to exceed the positive charge at the grid because it is the positive grid voltage that causes the plate-voltage drop. So the plate-voltage change counteracts only PART of the effect of the capacitor discharge, slowing the discharge considerably. This negative feedback keeps the capacitor discharge in the most linear part of the exponential curve. In fact, it can be shown mathematically that the time constant is lengthened by an amount equal to the amplification factor of this triode. Looking at it in still another way, the exponential curvature at the grid is counteracted by opposite curvature at the plate due to the inverting properties of the triode tube.

The discharge continues at a linear rate, as shown by the plate-voltage curve illustrated in figure 4–22. The increase in current raises the cathode voltage to 15.5 volts, and the grid 1 voltage is raised to plus 12 volts before the second balance point is reached.

Remember that during all this time, the screen grids (2 and 4) have been continually drawing current at an increasing rate. As the plate voltage continues to fall, there is a minimum point where the screen's current has risen sufficiently and the plate voltage has dropped to the point where the cathode potential is now high enough, in relation to the fixed potential on grid 3, to prevent any further increase in plate current. This occurs at about 108 volts on the plate. This point is also the peak of the plate-current characteristic (fig. 4–22). To the right of this peak, the plate current decreases and the screen current continues to increase. The screen grid's increasing current raises the cathode bias, thus the plate current stops increasing and levels off. This leveling off initiates the rapid switchover to recover the circuit to pretrigger condition.

When the plate current levels off, the voltage stops decreasing at the plate. With the counteracting effect of the plate drop removed, the capacitor discharge raises the grid 1 voltage at a very rapid rate. The positive-going grid 1 increases the current from the cathode, which raises the cathode voltage. This has the same effect as making grid 3 more negative; so grid 3 reduces the current to the plate. Actually, it is dividing the increasing current in favor of the screen grids. The screen current increases tremendously while the plate current is actually decreased. The plate voltage starts to go in the positive direction. This is coupled to the grid by the 500-$\mu\mu$f. capacitor to make it go in a positive direction. The cathode voltage is raised some more, which brings grid 3 closer to cutoff, reducing the plate current more and raising the plate voltage. This regenerative action continues until grid 3 is beyond cutoff—which stops plate current completely. During this regenerative action, the grid voltage has increased as fast as the plate voltage, so no change in voltage occurs. But when the grid reaches plus 40 volts, the plate current reaches zero and the grid no longer affects the plate voltage. Thus, as the grid voltage jumps from 12 to 40, the plate-voltage curve shows a 28-volt jump from 108 to 136 volts. When the grid voltage reaches a constant value, the plate rises only as fast as the capacitor can recharge along a slow exponential curve up to plus 300 volts. The cathode voltage rises immediately with the grid voltage while the screen grid voltage drops at recovery time because of the increased screen current.

Now consider the circuit from the viewpoint that it is ready for the next trigger pulse. The duration of operation, and consequently, the duration of the negative pulse on the cathode, is fixed in the circuit shown. This pulse width is determined essentially by the R–C in the grid 1 circuit, the plate voltage, and the gain of the triodes involved.

CHANGING PULSE WIDTH.—The duration of the pulse is generally made variable by applying a d-c voltage to the plate from a potentiometer. The circuit is as shown in figure 4–23. The potentiometer can select any voltage from zero to plus 300 volts. A diode is inserted in series to disconnect the phantastron plate from the low-resistance potentiometer during the operation of the circuit. Before the trigger pulse, the plate voltage and voltage at the arm of the potentiometer are the same because the load resistor is so large and the current from the potentiometer so small that there is little voltage drop in the diode plate. When the circuit operates, the plate voltage drops immediately, the diode plate becomes less positive than its cathode, and diode current stops. The potentiometer-diode circuit has no further effect on the operation of the phantastron.

To see how the potentiometer affects the pulse width, study the effect of varying the position of the potentiometer contact. With the plate voltage at plus 300 volts, the situation is as previously described. But when this plate voltage is reduced to plus 240 volts by means of the potentiometer, the pulse width is decreased. The plate voltage drops the same 42 volts at the time of the trigger pulse since the grid drops 42 volts to the first balance point. Then the slow linear voltage decrease occurs. The slope of this decrease is the SAME regardless of plate voltage. This terminates at the same 108 volts as the 300-volt waveshape. So the amount of linear decrease is less, which decreases the time required to reach 108 volts. If the potentiometer is set so the plate voltage is 170 volts before the pulse, the plate will again drop 42

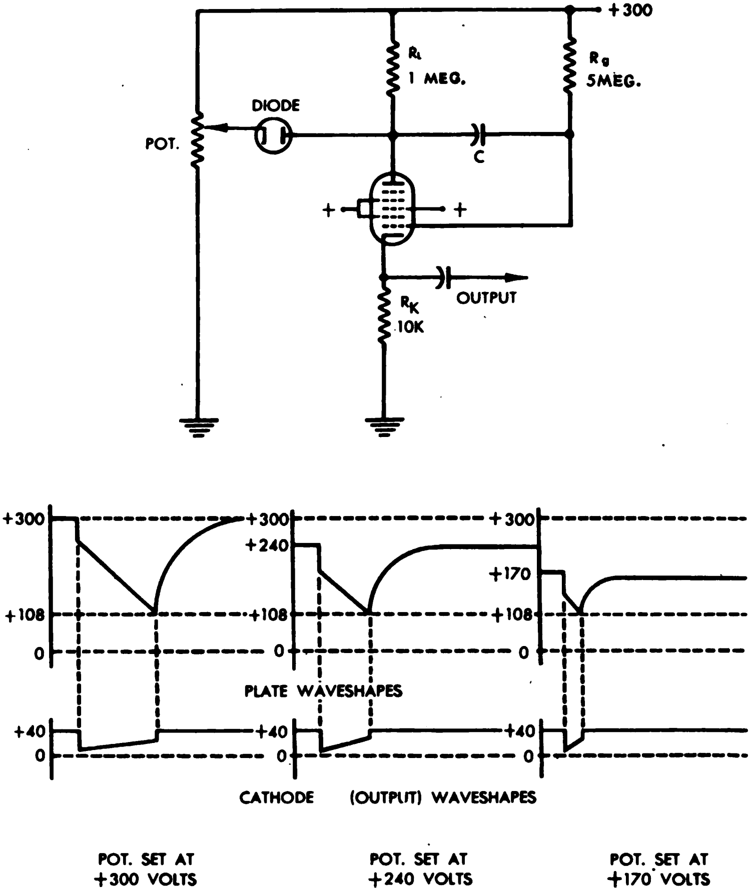


Figure 4–23.—Changing pulse width in phantastron circuit.

volts, and follow the SAME slope to 108 volts. Again the amplitude of the slope is decreased so the slope time is decreased.

Since the initial drop, the slope, and the final value do not change, the decrease in pulse length is linear with the decrease in plate voltage. This relation holds over the range from the full plate voltage (300 volts in this case) to a minimum of 108 plus 42, or 150 volts. This is the minimum voltage point because the initial 42-volt drop would bring it down to 108 volts, or zero pulse width. Since the pulse width for the previous example used is 1,395 microseconds, the pulse width is variable from zero to 1,395 microseconds.

APPLICATION OF PHANTASTRON CIRCUIT.—The phantastron circuit shown in figure 4–24 is taken from a typical radar set. It is similar to the previous circuit except that there the values and voltages are slightly different, the voltage dividers for grids 2, 3, and 4 are shown, and the disconnector diode is actually a triode with its grid tied to the plate. This last is done in the set to make use of the other half of a 6SN7 that is employed in another circuit. The potentiometer which sets the plate voltage has fixed resistors in each end to limit the pulse width to a certain maximum and minimum width which is adequate for the purpose for which it is designed.

The trigger pulse is applied to the cathode of the disconnector diode. The pulse momentarily changes the plate current in the diode, which causes the negative pulse to appear at the plate of the phantastron tube. The plate-to-grid coupling capacitor couples the pulse voltage to the grid where it starts the circuit through a cycle of operation. The output is taken from the cathode of the 6SA7, where a negative, approximately square waveshape is produced. Actually, the waveshape is not as ideally square as shown in figure 4–22, because the trailing edge becomes somewhat exponential in following the capacitor charge.

In use, the square wave is usually differentiated and the trailing edge pulse is put through a pulse sharpening circuit to remove the effect of the exponential trailing edge. The resulting sharp pulse is delayed from the original trigger pulse by a few hundred microseconds.

PHANTASTRON WITH CATHODE FOLLOWER.—The previous circuit is satisfactory for short delays, but when the delay becomes nearly as long as the time between pulses, the long exponential recovery of the plate circuit overlaps the next pulse. With the circuit not yet fully recovered,

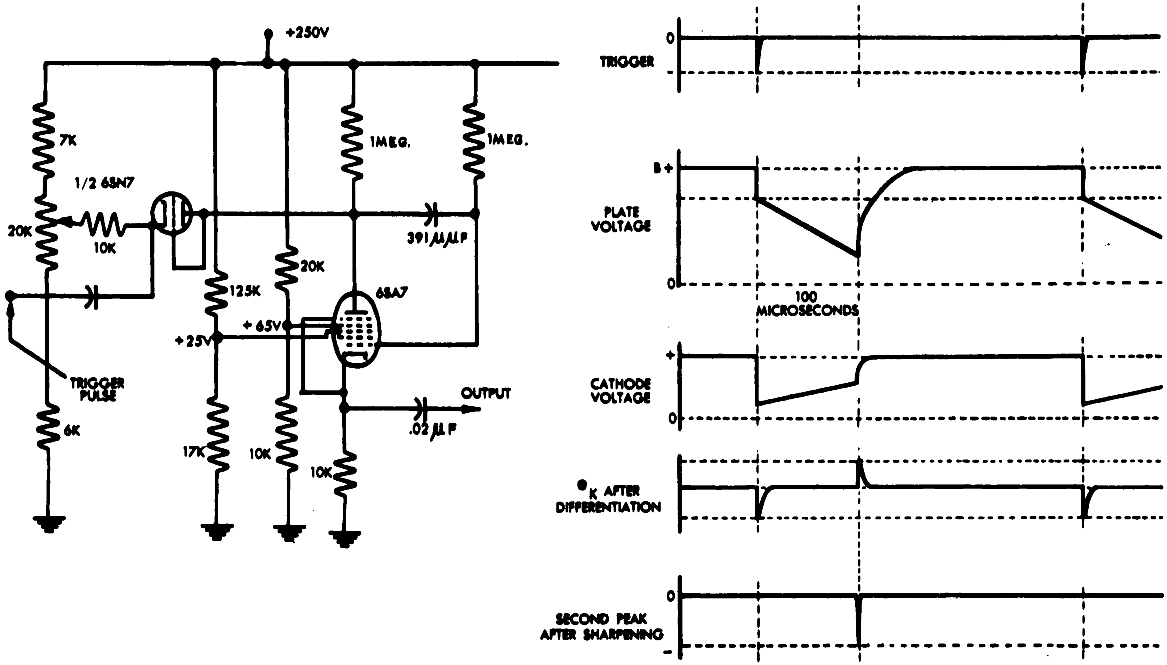


Figure 4–24.—Typical radar phantastron circuit.

the second pulse may not trigger the circuit. Therefore the circuit is modified for long delay periods as shown in figure 4–25. The change is the addition of a triode between the plate and grid of the phantastron tube. This triode is called a CATHODE FOLLOWER. Before the trigger pulse arrives, the grid of the cathode follower is at the plate potential, it is drawing current, and the cathode is within a fraction of a volt of the grid voltage. Therefore the 0.005-μf. coupling capacitor has the plate voltage on one side and the grid voltage on the other—which is not different from the circuit in figure 4–22. When the trigger pulse drops the plate voltage, the grid voltage of the cathode follower, and consequently its cathode voltage, goes down. The 0.005-μf. capacitor must discharge, and in doing so, makes grid 1 go in the negative direction. Thus by cathode follower action, the grid drops right along with the plate as though capacitive coupling existed between them.

When the end of the pulse arrives, and the plate voltage rises back to 250 volts, it does not have to charge the capacitor. The plate can raise the cathode follower grid immediately to plus 250 volts. Then the capacitor charges through a path from the 6SA7 cathode to grid 1, through the capacitor, from cathode to plate of the cathode follower, to plus 250 volts. The grid of the cathode follower is not involved, other than through interelectrode capacity, which is very small. Only a slight exponential curvature of the plate waveshape is shown in figure 4–25. This curve is due to stray wiring capacity in the circuit. The dotted curve shows the exponential rise which would interfere with the next pulse if no cathode follower were used. In all other respects, this long delay circuit is the same as the short delay circuit. In some equipments the square wave output is taken from the grids 2 and 4. This is a positive square wave, and when differentiated produces a negative pulse at the square wave trailing edge.

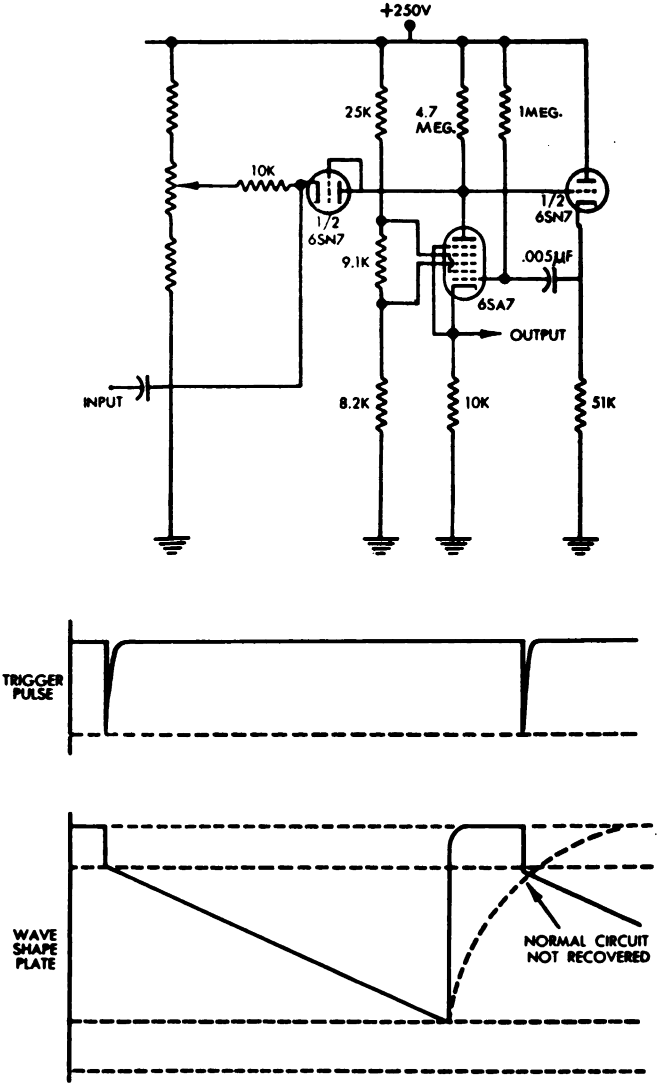


Figure 4–25.—Phantastron with cathode follower.

THE BLOCKING OSCILLATOR

The blocking oscillator is another type of relaxation oscillator. It can do many of the things that a multivibrator does. Thus it can be free running, synchronized at each cycle or at a submultiple frequency, or it can be driven in the manner of a one-shot multivibrator.

A blocking oscillator is any oscillator which cuts itself off after one or more cycles because of the accumulation of a negative charge on the grid capacitor. Thus, in an oscillator in which the grid swings positive with respect to the cathode, electrons are attracted to the grid and accumulate on the plate of the grid capacitor nearest the grid. Since these electrons cannot return to the cathode through the tube, they must return through the grid-to-cathode resistor. If this resistor is sufficiently large, electrons may accumulate on the capacitor faster than the resistor permits them to return to the cathode. In this case a negative charge is built up at the grid which may bias the tube beyond cutoff. After the tube is cut off, it provides no additional electrons to the grid capacitor. However, the capacitor continues to discharge through the resistor, and a point is reached eventually where the tube again conducts. Thus the process is repeated and the tube becomes an intermittent oscillator. The rate of the recurrence of operating conditions is determined by the R–C time constant of the grid circuit.

There are two general types of blocking oscillators—the SINGLE-SWING type in which the tube is cut off at the completion of one cycle, and the SELF-PULSING type, in which each cycle of oscillation causes the grid to become progressively more negative until the tube is biased out of operation. In radar the single-swing type oscillator usually operates within the audiofrequency range, while the self-pulsing type produces pulses of RF energy.

Single-Swing Blocking Oscillator

The single-swing oscillator circuit shown in figure 4–26 consists of a transformer-coupled oscillator, with a capacitor in series with the grid of the triode V_1. For

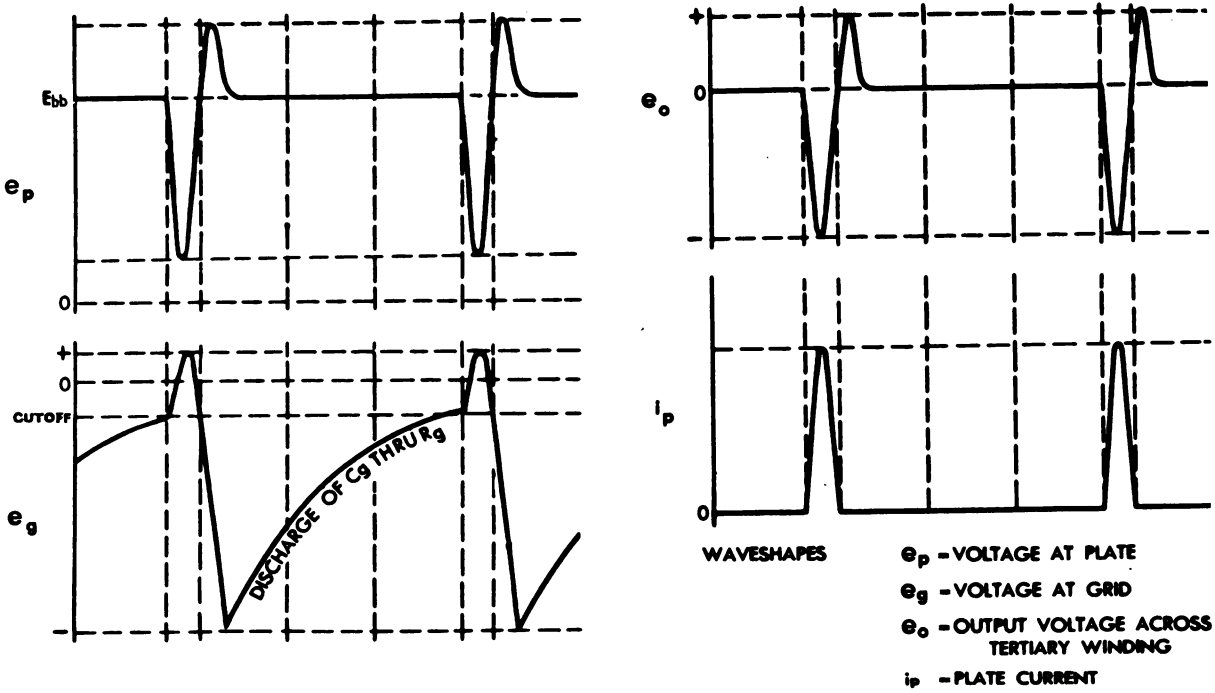
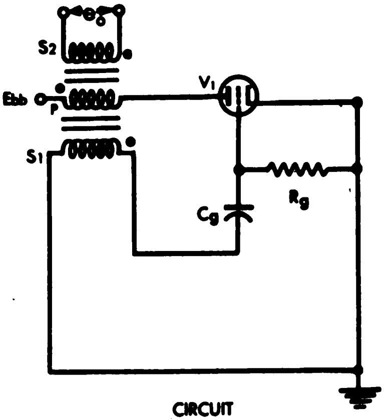
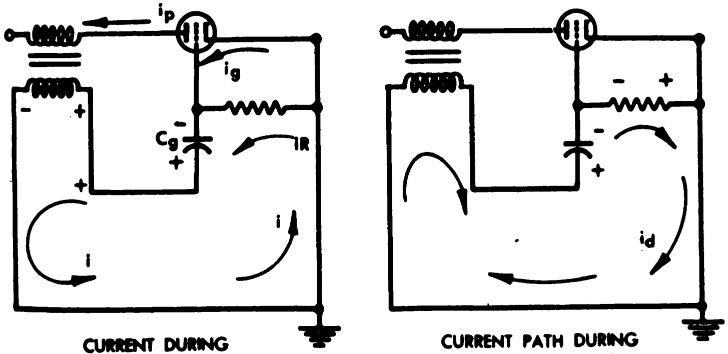


Figure 4–26.—Single-swing blocking oscillator.



CIRCUIT



CURRENT DURING CHARGE OF Cg

CURRENT PATH DURING DISCHARGE OF C_g

Figure 4–26.—Single-swing blocking oscillator. Continued

the purpose of understanding its operation, assume that the grid capacitor C_g has been negatively charged by a preceding cycle. The tube, therefore, is biased well below cutoff. As the charge on the capacitor leaks off, the biasing voltage is reduced to the point where the tube begins to conduct. As plate current starts to flow, a magnetic field is set up around the plate winding *P* of the transformer. The dots at each winding indicate similar polarities. If, for example, a current flows through a winding so that the dot end is positive, the field set up in the core induces voltages in the other windings that make the dot end positive in these windings at the same time.

The field builds from zero to a maximum in direct proportion to the plate current, and therefore induces a voltage across the grid winding. This voltage is of such polarity that the grid end goes positive. This action is impressed upon the grid of the tube through the grid capacitor C_g with a polarity that drives the grid more and more positive as the field in the plate winding is building up. When the grid is driven positive with respect to its cathode, it draws current, and electrons accumulate on the plate of the grid capacitor nearest the grid. As the grid becomes more positive, it eventually takes enough electrons from the plate current to lower the rate of the plate current increase. This reduces the induced voltage, the grid stops swinging positive, and the plate current increase is slowed. This action is cumulative and ends with the plate current being settled at a constant value, and the magnetic field becoming stationary with respect to the secondary.

At this time there is no induced voltage in the grid winding and, because there is no charging potential applied, the capacitor begins to discharge. This discharge causes the potential on the grid to become less positive, thereby causing less plate current to flow in the plate winding. The field around the plate coil starts to collapse. This collapsing field, in turn, induces a voltage in the grid winding in the reverse direction, causing the grid to become more and more negative. This process continues until the grid is driven beyond cut-off, thus completing a cycle of operation. Oscillation does not start again immediately, however, because the flow of grid current when the grid was positive has built up enough charge on the grid capacitor to hold the tube cut off until some of the charge leaks off through the grid resistor.

The time consumed by the rise and decay of plate current depends upon the inductance and resistance of the transformer. The time between pulses depends primarily upon the value of resistance since the grid capacitance is fixed because of pulse-width requirements.

Note the current paths during charge and discharge. The capacitor charges very quickly through the low resistance of the conducting grid-cathode circuit but discharges very slowly through the large grid-leak resistance. The rounded waveshapes show that the voltage and current changes occur more slowly than in a multivibrator. This is due to the inductance in the transformer. The inductance also is the cause of the plate voltage rising above E_{bb} when the tube is cut off. The current continues to flow in the same direction as it charges the distributed capacity in the transformer windings, and causes a voltage drop which adds to E_{bb}. The inductance and distributed capacity form a resonant circuit which tends to oscillate due to shock excitation. If the Q of this L–C circuit is high, it may oscillate for several cycles. In the case illustrated (fig. 4–26), damping is at the critical value so that the oscillation occurs for about a half cycle.

The ouput e_o from this circuit is taken from a third winding on the transformer. However, the output can also be taken from the plate, where a large negative pulse is produced, or from a small resistor in the cathode circuit, where a small but sharp positive pulse is available. The waveshape of the cathode pulse is the same as the plate current waveshape shown in figure 4–26.

Synchronized Blocking Oscillator

The synchronized blocking oscillator in figure 4–27, is taken from a radar set. This oscillator is used as the timing oscillator in a small piece of auxiliary equipment. Normally, the set produces its own PRF independently when the oscillator is free running, as you can see in the waveshapes labeled "e_g, no sync." The natural frequency for its operation is about 300 c.p.s. However, synchronizing pulses can be introduced at the plate without making changes in its circuit that may vary the PRF. The synchronization pulses shown occur at 700 c.p.s. Their amplitude is such that one pulse makes the circuit oper-

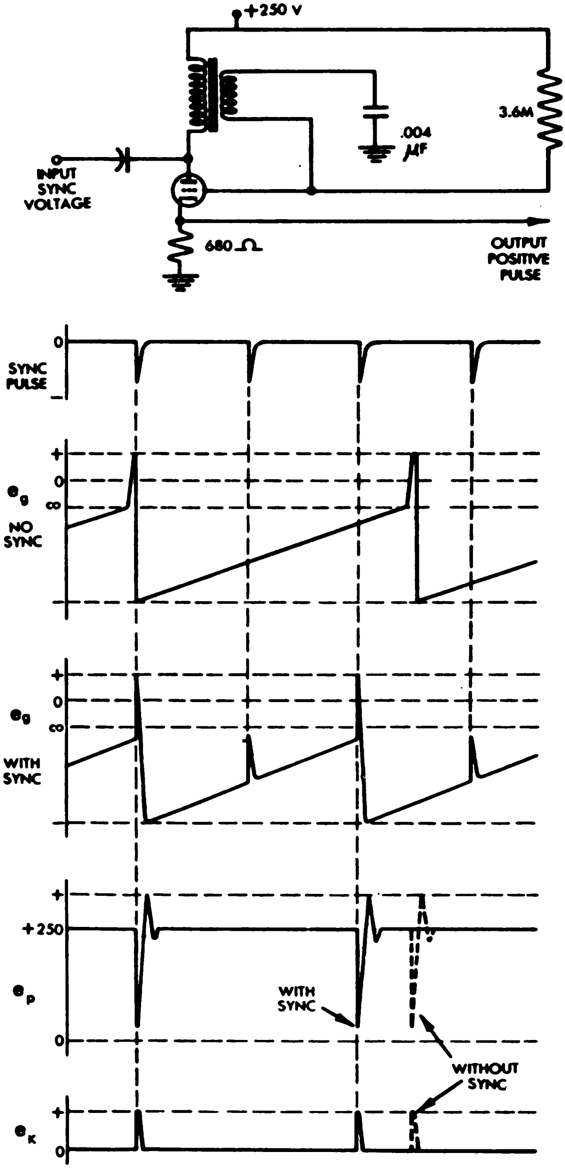


Figure 4–27.—Synchronized blocking oscillator.

ate, the next pulse does not bring the grid above cutoff, but the third pulse does make the tube conduct. Since the oscillator is synchronized on alternate pulses, its PRF becomes 700/2, or 350 c.p.s. If the main radar set has a PRF of 1,000 c.p.s. and the amplitude slightly less than that shown, the blocking oscillator can be synchronized on every third cycle, and it will operate at 333⅓ c.p.s. By synchronizing this auxiliary set with the big set, indications received by the small set can be displayed on the main indicator, relieving the operator of observing two radar indicators at the same time.

The amplitude of the synchronizing pulse has a great effect on the frequency of this circuit. If the amplitude of the synchronization pulse shown is increased 50 percent, the first pulse that occurs after the circuit begins operation will bring the grid above cutoff and the circuit will be synchronized on every pulse. As stated before, a low amplitude could cause triggering on every third pulse since the first two will not have sufficient amplitude to make the tube conduct.

In each case, it is assumed that the natural period of of the circuit is longer than the time between the pulses which cause the circuit to operate. If the period of the blocking oscillator is shorter, it will produce long and short cycles because the circuit will be triggered on some pulses but will conduct on its own accord at times.

Maximum frequency stability is achieved in this circuit by connecting the grid-load resistor to the plate-voltage supply. This has the same effect as returning the grids to B plus in multivibrators. The useful output from this circuit is taken from a small (680-ohm) resistor in the cathode circuit. This resistor has a negligible effect on the operation of the circuit but develops a voltage pulse due to the plate current flowing through it.

The Driven Blocking Oscillator

Some electronic circuits require a blocking oscillator which operates only when triggered. Such a circuit might be called a BLOCKED OSCILLATOR since a permanent d-c

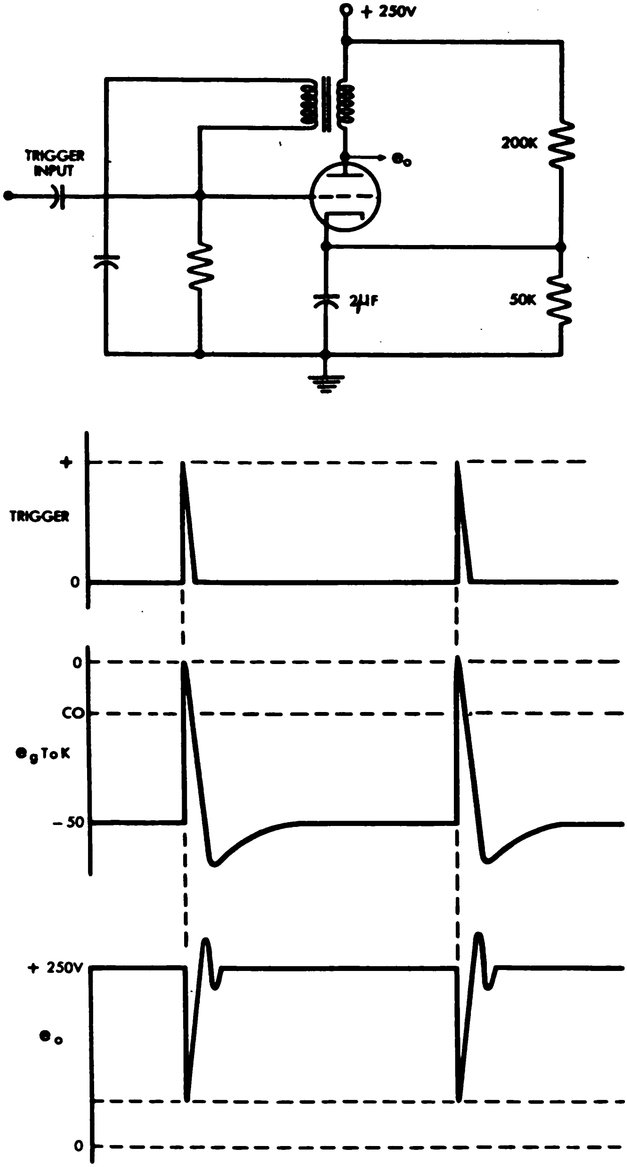


Figure 4–28.—Blocked oscillator.

voltage prevents the circuit from oscillating of its own accord. In the circuit of figure 4–28, only a positive voltage on the grid or a negative voltage on the cathode will cause the tube to conduct. If the voltage is a pulse, the circuit will go through a cycle of operation, cut itself off, and the d-c voltage will keep it cut off. Cutoff is maintained by a voltage divider from the plate-supply voltage. With the values shown, the cathode voltage will be plus 50 volts with respect to ground. The grid is grounded, and is therefore 50 volts negative with respect to the cathode. The grid waveshape shows voltages with respect to the cathode since these are the important voltages in understanding the circuit operation. Cutoff for most tubes used in this type circuit occurs at about minus 18 volts, so the trigger pulse must be more than 50 minus 18, or 32 volts in amplitude to start a cycle of operation. In the grid waveshape the normal discharge of the grid capacitor reduces the grid voltage after a cycle of operation, but the discharge is to a minus 50 volts, so the circuit does not again conduct without another trigger pulse.

The positive trigger pulse must be applied to the grid only. It is not practical to apply a trigger pulse to the cathode because the filter capacitor is usually found there. But a negative pulse may be applied to the plate, where the transformer will invert the pulse to a positive one on the grid. The output may be taken from the plate, where a large negative pulse is available, or from a third transformer winding where a pulse similar to the plate pulse can be obtained with either polarity.

USE AS A PULSE SHARPENING CIRCUIT.—The driven or blocked oscillator is useful as a pulse sharpening circuit. One set uses the circuit in figure 4–29 to improve the pulse shape from the phantastron. The phantastron produces the square wave labeled input waveshape. This square wave is amplified and becomes the input to the differentiating transformer shown in the circuit. The waveshape at the secondary of the transformer is a

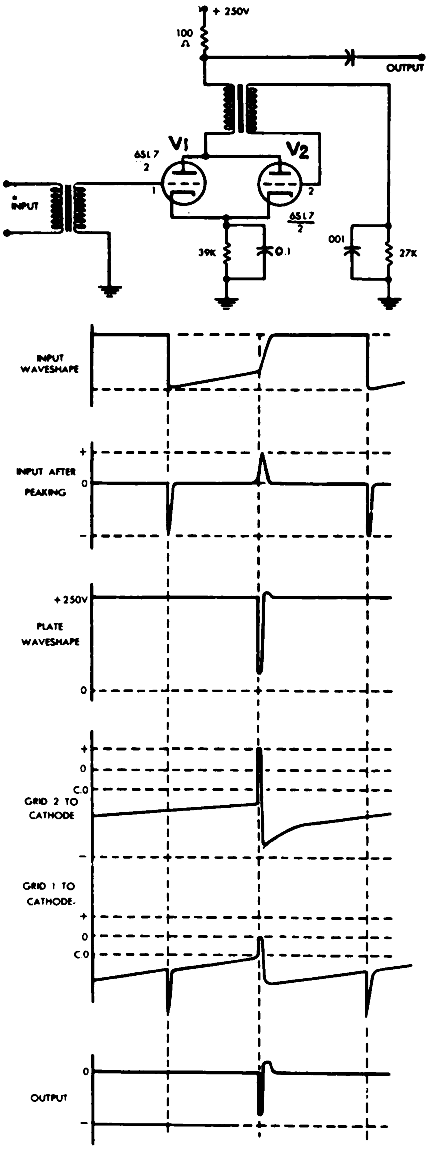


Figure 4–29.—Blocked oscillator as a pulse sharpening circuit.

peaked wave, as shown in the second waveshape, but because of the inherent imperfections in the trailing edge of the phantastron pulse, the peaked wave due to the trailing edge is not a sharp one. For precise triggering of the circuits following the phantastron, a sharper pulse is required. Therefore the peaked wave is applied to the blocked oscillator.

In place of the voltage divider from plus 250 volts used in the earlier circuits, a long time constant in the cathode circuit provides the cutoff voltage for the preceding circuit. After several cycles of operation, the grid current due to positive grid voltages charges the cathode capacitor. It will discharge slowly between cycles and be recharged with each cycle. The R–C combination in the grid circuit is rather short and cannot hold the grid below cutoff for the several hundred microseconds between pulses.

The negative pulse at the leading edge of the square pulse will not affect the already cutoff tube. The positive pulse at the trailing edge will cause the left tube to conduct. The current through the transformer and associated magnetic field will induce a positive voltage at the grid of tube V_2. This starts a regenerative action in the blocking oscillator associated with tube V_2 and it goes through a normal cycle of operation. Meanwhile the pulse ends at the grid tube V_1, and with no other voltage on the grid, the tube is cut off again.

Tube V_1 is an amplifier and isolation circuit which causes the blocking oscillator to start very soon after the trigger pulse arrives, then disconnects the triggering circuit from the blocking oscillator so it can go through its normal cycle unaffected by the triggering circuit.

The output is taken from a small resistor above the transformer in the plate lead. A sharp negative pulse of low amplitude is produced across the resistor.

COUNTING CIRCUITS

A counting circuit, also known as a frequency divider, is one which receives uniform pulses, representing units to be counted, and produces a voltage proportional to their frequency. By slight modifications the counting circuit is used in conjunction with a blocking oscillator to produce a trigger pulse which is a submultiple of the frequency of the pulses applied.

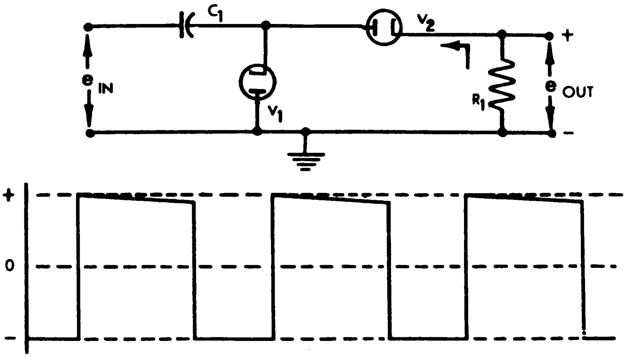
The pulses applied to the counting circuit must have the same amplitude and time duration if accurate frequency division is to be made. Counting circuits, therefore, are ordinarily preceded by shaping and limiting circuits to insure this uniformity of amplitude and width. Under these circumstances the pulse-repetition frequency constitutes the only variable, and frequency variations may be measured.

The Positive Counter

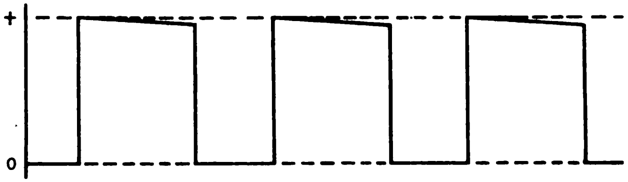
Positive pulses, which may vary only in their recurrence frequency, are applied to the input of the positive counter as shown in figure 4–30. The charge on the coupling capacitor C_1 cannot change instantaneously as the positive leading edge is applied, so the plate of V_2 becomes positive and the diode conducts. A charging current flows through R_1 during the pulse time and a small charge is developed on C_1. At the end of the pulse the drop in voltage places the diode side of the capacitor at a negative potential equal to the charge accumulated on C_1.

Tube V_2 cannot conduct, as its plate is negative with respect to its cathode. However, V_1 conducts, discharging the small charge from the capacitor, which would otherwise build up during each succeeding positive pulse, eventually rendering the circuit insensitive to the applied pulses.

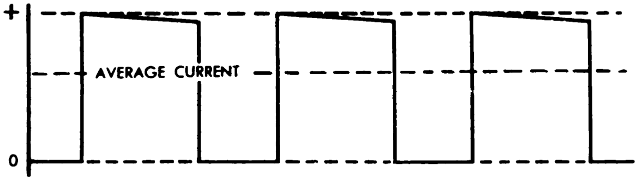
CIRCUIT CONTROLLED BY POSITIVE COUNTER.—Since a certain amount of current flows through R_1 each time a pulse is applied, an average current flows which increases as the pulse recurrence frequency increases and decreases



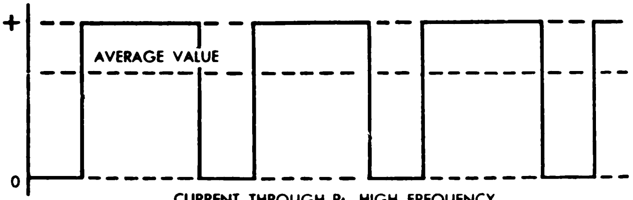
e_{IN} (INPUT VOLTAGE)



VOLTAGE AT PLATE OF V_2



CURRENT THROUGH R_1, LOW FREQUENCY



CURRENT THROUGH R_1, HIGH FREQUENCY

Figure 4–30.—The positive counter.

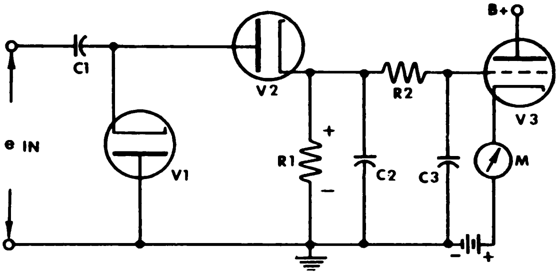


Figure 4–31.—Circuit controlled by positive counter.

as this frequency decreases. The *IR* drop developed across R_1 can be used to control a succeeding stage as is illustrated in the circuit controlled by a positive counter (fig. 4–31). The filter in the grid circuit of V_3 aids in obtaining smooth operation by removing too rapid changes in voltage developed across R_1. The voltage at the grid of V_3 varies with changes in the pulse frequency and produces variations in the plate current by V_3. A milliammeter is placed in series with the plate circuit so that changes in the average plate current are indicated as a measure of variations in the recurrence frequency of the input pulses. Frequent use of this counter circuit can be found in radio altimeters.

The Negative Counter

By reversing the connections to the diodes V_1 and V_2 in the positive counting circuit the device can be made to respond to negative pulses, and thus it becomes a negative counter circuit. This circuit is shown in figure 4–32. The diode conducts during the time the negative pulse is applied and an electron current flows through R_1 as indicated by the arrow. At the end of the negative pulse V_1 conducts sufficiently to remove the small charge which developed on C_1 during the pulse time. The current through R_1 increases with an increase in the pulse frequency as before. However, if the voltage developed across R_1 is applied to the same control tube shown previously, the increase in current causes the grid of V_3 to become more negative. This decreases the plate current

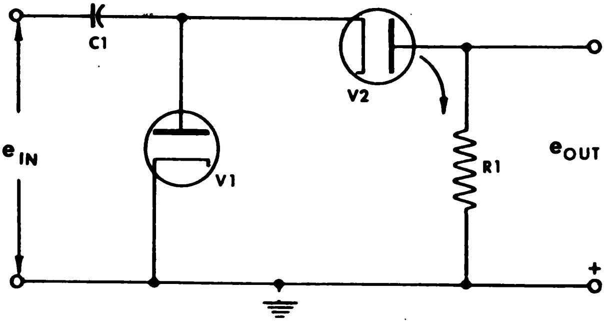


Figure 4–32.—Negative counting circuit.

through V_3 and the meter. Thus an increase in the frequency of the negative pulse causes a drop in the average plate current measured by the meter. This effect is the opposite to that in the positive counter.

The Step-by-Step Counter

The step-by-step counting circuit is similar to those already discussed except that a capacitor which is large compared to C_1 replaces the resistor R_1 used in tne positive counting circuit. The charge on this capacitor, C_2 (fig. 4–33) increases slightly during the time of each positive pulse and produces a step voltage across the output. These steps decrease in size exponentially as the voltage across C_2 approaches the final value, the rate being dependent upon the output impedance of the driving circuit. As long as there is no path through which C_2 can discharge, the voltage across it continues to increase with each successive pulse until it is equal to the amplitude of the applied signals. At this point the cathode of V_2 is held at a positive voltage equal to that on the plate during the pulse time and the tube fails to conduct.

CIRCUIT ACTION AND WAVESHAPE ANALYSIS.—Use is made of the step charging circuit to trigger a blocked oscillator after a certain number of steps. A simple circuit of this type is shown in figure 4–34. The step-by-step counter determines the voltage on the blocked oscillator grid. The cathode has a high fixed bias on it, and

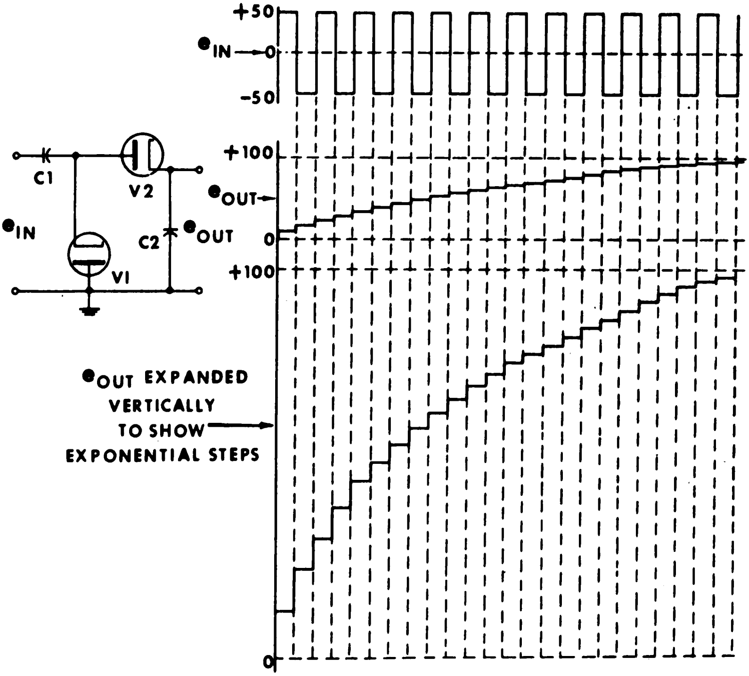


Figure 4–33.—Step-by-step counting.

a certain number of steps are required to raise the grid near enough the bias level to make the tube conduct. When the tube conducts, the blocking oscillator circuit goes through a normal cycle. The usual grid current during this cycle discharges the capacitor C_2. At the end of the cycle, C_2 charges in steps again, starting from zero.

This circuit is not difficult to analyze quantitatively. The following is the calculation of the exact frequency division ratio in the circuit. In the step-charging circuit, the time constant of the charging circuit is short since only the diode plate resistance is in series with the capacitors. Therefore, the capacitance is fully charged during each cycle of input waveshape. But the capacitors are in series, so the full charging voltage will be divided between the two. The voltage distribution in capacitors is in inverse ratio to the values of the respective series-connected capacitors and is mathematically expressed,

$$\frac{C_1}{C_2} = \frac{E_2}{E_1}$$

In this equation,

C_1 = Capacitance of first series capacitor.

C_2 = Capacitance of second series capacitor.

E_2 = Charge on C_2 if applied voltage is E_1 plus E_2.

E_1 = Charge on C_1 if applied voltage is E_1 plus E_2.

The voltage in question is E_2. Let the applied voltage be E_a. The equation for the E_2 voltage becomes:

$$E_2 = E_a \frac{C_1}{C_1 + C_2}$$

Using the values indicated in figure 4–34,

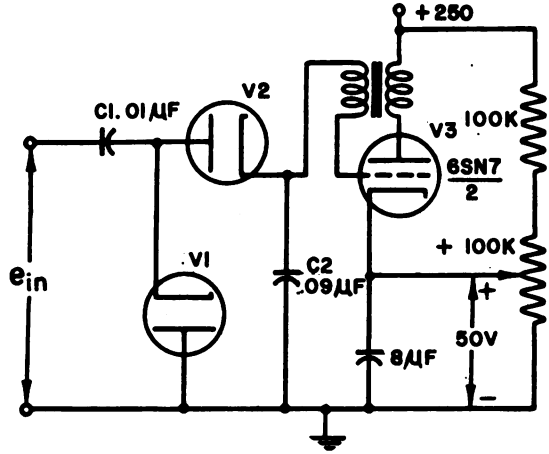
$$E_2 = 100 \times \frac{0.01}{0.1 + 0.09}$$

$$E_2 = 100 \times \frac{0.01}{0.1}$$

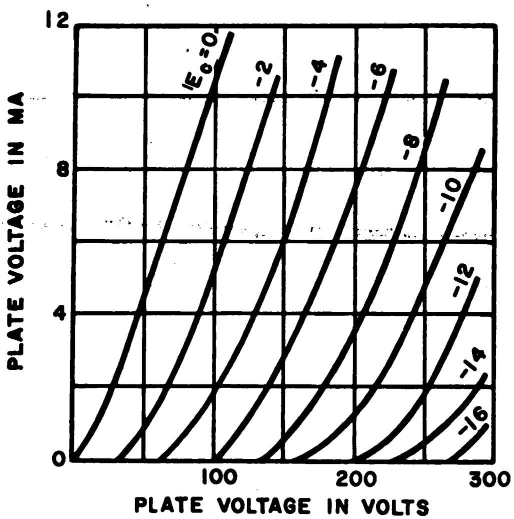
$$E_2 = 10 \text{ volts}$$

The charge on capacitor C_2 with 100 volts applied will be 10 volts, while the capacitor C_1 will charge to 90 volts. It follows that the first step is 10 volts, which is 10 percent of the applied voltage. The first step is 10 volts because the input waveshape is 100 volts peak-to-peak and the first capacitor and diode V_1 form a DCR (direct-current restorer) circuit which allows only positive voltages on the plate of the second diode, V_2. Therefore the applied voltage is plus 100 volts during the pulse and C_2 is charged to 10 percent of the total.

The first input pulse ends, C_1 is discharged, but C_2 holds its 10-volt charge. The second pulse occurs. The 10-volt charge on C_2 will oppose the 100 volts of the second pulse, and the total applied voltage, to which the



CIRCUIT



6SN7 CHARACTERISTICS

Figure 4–34.—Triggering blocked oscillator with step-charging circuit.

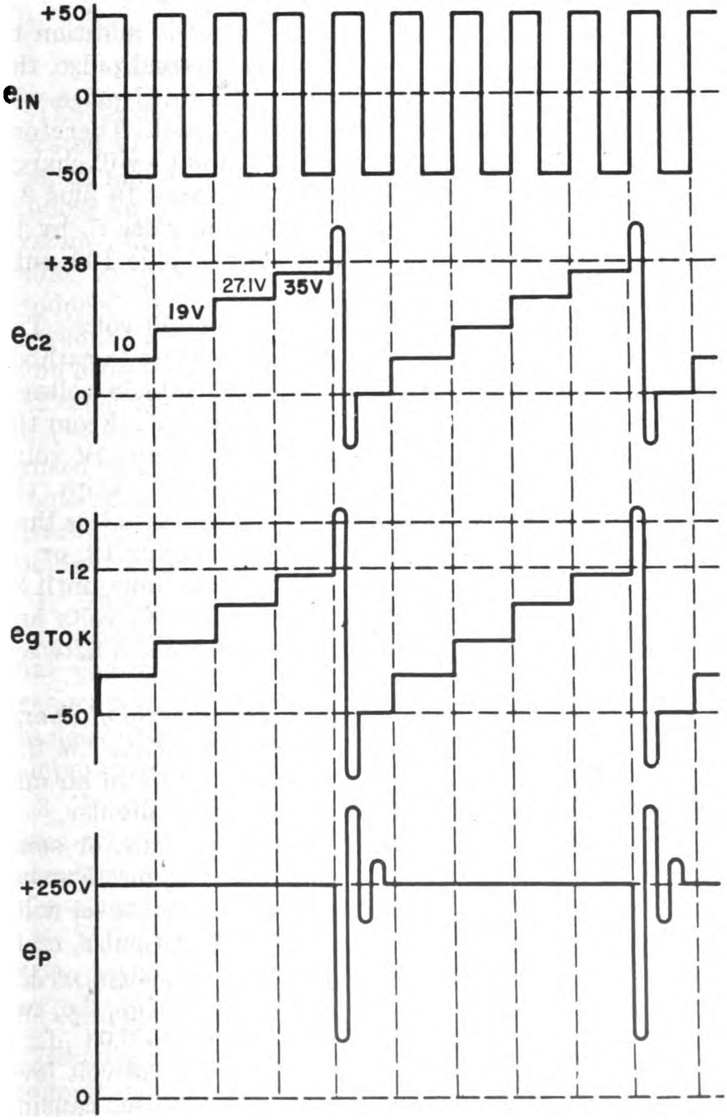


Figure 4–34.—Triggering blocked oscillator with step-charging circuit. Continued.

capacitors can charge is then 90 volts. E_a will be 90 volts, and from the previous equation, C_2 will again be charged by 10 percent or 9 volts. This is in addition to the initial 10 volts, so at the end of the second pulse, the C_2 voltage is 10 plus 9 or 19 volts. The third pulse will be 100 volts again, but 19 volts will oppose it. Therefore the E_a will be 100 minus 19 or 81 volts and C_2 will charge an additional 8.1 volts. The new E_2 becomes 19 plus 8.1 or 27.1 volts. Successive input pulses will raise C_2 by 10 percent of the remaining voltage toward plus 100 until the blocked oscillator is actuated.

In the oscillator, the bias is set at plus 50 volts. The plate-supply voltage is 250 volts, so the plate-to-cathode voltage is 200 volts. Using the plate-to-cathode voltage, you can determine the grid voltage for cutoff. From the curves in figure 4–34, this value is near minus 12 volts, as indicated on the characteristic curves. With the cathode at plus 50 volts, the grid must be within less than 12 volts of this value, that is, above 50 minus 12, or 38 volts to allow conduction. So the steps continue until 38 volts is exceeded. Since the fourth step is 35 volts and the fifth step 41 volts, the 38-volt level is crossed between the fourth and fifth steps.

The blocked oscillator goes through one cycle of operation at the end of which the "bucket capacitor" is discharged. Since every fifth input pulse results in an output pulse, this circuit is a 5-to-1 frequency divider.

To cause the circuit to divide by three, four, or some other value, the bias is set at some new value. For example, a bias of 28 volts would cause the crossover with cutoff voltage to occur at the end of the third pulse, making the circuit a 3-to-1 count divider. The sizes of the steps can be varied by changing the ratio of the two capacitors. For example, if C_2 is changed to 0.04 μf., it will take 20 percent of the charge, and the 38-volt level will be crossed at the end of the third pulse, causing division at a 3-to-1 ratio. In practice, this frequency division can be carried to the extent that with six successive frequency dividers, a 100,000 c.p.s. sine wave can be reduced to 50 c.p.s. The final circuit will be triggered on each two-thousandth cycle of the original voltage.

AUTOMATIC FREQUENCY CONTROL (AFC)

The term AFC refers to an automatic system incorporated in a radio or radar receiver circuit for the purpose of maintaining the correct local-oscillator frequency. Almost all Navy radar equipments make use of an AFC system. In most cases, receiver equipments, other than radar, are tuned to a fixed frequency. This frequency is produced by a transmitter that is capable of accurate frequency control and stability, and because of this there is no need for an AFC system.

Transmitters that produce frequencies in the UHF and microwave ranges have poor frequency stability. This is caused by factors which have negligible effect at lower frequencies. A partial listing of the cause of frequency instability at UHF and microwave ranges in transmitter oscillators and local oscillators is as follows:

1. Small physical size. Since the frequency determining components are small in physical size any condition that changes their shape or size will cause a shift in frequency. Factors that cause these components to change their characteristics are vibrations, variations in temperature, and variations in atmospheric pressure.

2. Frequency pulling. This is the result of load variations which are caused by variations in the length of transmission lines or reflections from nearby objects, such as sea return, aircraft surfaces, or the radome.

3. Voltage variations. These are caused either by variations in the source of power or poor regulation in the equipment's power supply.

Because of the previously presented inherent conditions it is important that the tuning of a microwave receiver to the proper frequency be automatic.

Some of the properties of microwave oscillators which may be used for frequency control are:

Those of a magnetron—mechanical, thermal, pulling, and varying flux field density.

Those of a klystron—mechanical, thermal, pulling, and electrical.

The system most often used and the one that will be described for the purpose of enabling you to gain a fuller understanding of AFC is the system which controls the repeller voltage of the local oscillator. This is the more practical method because a change in frequency is nearly linear with a change in voltage; also, changes up to 60 megacycles are practical. The operation of this type of oscillator, known as the klystron, will be explained in detail in chapter 6.

CLASSIFICATION OF AFC SYSTEMS.—AFC systems are classified in accordance with their method of operation. The method of operation may be divided into two major classifications, the ABSOLUTE and the DIFFERENCE FREQUENCY systems.

In the absolute frequency system the frequency of oscillation is determined by a precision frequency standard such as a high-*Q* cavity. This system is generally used to keep the receiver tuned to the beacon (RACON) frequency. In the difference frequency system the local oscillator is maintained on a frequency which is greater or less than the magnetron frequency by an amount equal to the intermediate frequency. This keeps the receiver tuned to the magnetron during search operation.

Either of these systems may be a nonhunting or hunting type construction. The hunting system is constructed to cause the oscillator to sweep a wide frequency range. This enables the oscillator to find the correct operating frequency and once this has been accomplished the oscillator will automatically remain on the correct frequency. This system has the advantage that in the event of a sudden shift in magnetron or local-oscillator frequency the correct operating frequency will be found automatically. In the nonhunting system the oscillator is manually tuned to the correct frequency and then maintained electronically to this frequency.

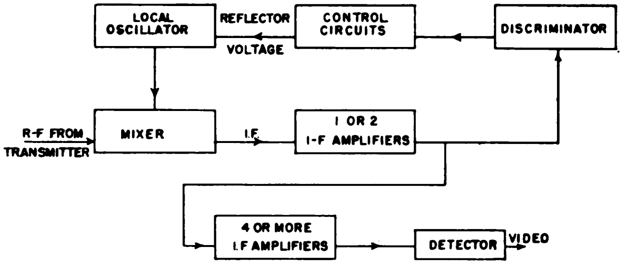


Figure 4–35.—Block diagram of AFC circuit in a radar receiver, single-mixer system.

BLOCK DIAGRAM OF AFC SYSTEM.—In the single-mixer system (fig. 4–35), a portion of the transmitter pulse leaks through the TR box and beats with the local-oscillator signal in the mixer. The difference frequency, or the IF signal, appears at the output of the mixer. This IF signal is then amplified by one or two stages of IF amplification. At this point the signal follows two paths—it is further amplified and fed into the detector, and it is also fed into the AFC discriminator. The output of the discriminator is fed to either a d-c amplifier or some special control circuits which regulate the klystron reflector voltage. The regulation of this voltage changes the frequency of the klystron and enables the local oscillator to oscillate at the correct frequency.

Most modern radar systems use a double-mixer AFC system, as shown in figure 4–36. The separate signal and the AFC mixer are both connected to the same local oscillator. This arrangement is shown in figure 4–36. The AFC mixer is coupled to the transmission line

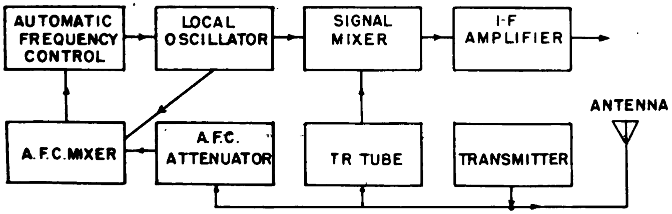


Figure 4–36.—Block diagram of a double-mixer AFC system.

through an attenuator which reduces the transmitted pulse and received signals by the same ratio. With this arrangement, strong signals from nearby radar, jammers, or sea return cannot affect the operation of the AFC loop. The double-mixer system is similar to the single-mixer system with the exception that it contains an AFC attenuator and separate IF stages.

THE DISCRIMINATOR CIRCUIT.—The amplified output of the AFC mixer is fed into the AFC discriminator. The output voltage of a frequency discriminator varies in amplitude and polarity with changes in the applied frequency. If the applied frequency is at the resonant or center frequency of the discriminator the output will be zero. If the applied frequency is above or below the correct value, the output of the discriminator will be either positive or negative, depending upon the direction of deviation. The operation of this type of discriminator is explained in detail in chapter 8, *Basic Electronics,* NavPers 10087.

THE THYRATRON AFC CONTROL CIRCUIT.—This circuit is an automatic hunting and lock-in type. A simplified diagram is shown in figure 4–37. Since the discriminator time constant is short, a pulsed output is produced. These pulses are amplified and inverted by the video pulse amplifier.

In the circuit shown in figure 4–37, V_2 functions as a sawtooth sweep generator. The output of this generator is applied to the repeller of the klystron. This causes a wide frequency band to be swept by the klystron and provides a "hunting action," which is a requirement of the AFC system. Lock-in action is obtained by V_1 once the proper IF signal is produced. During the time that V_1 maintains lock-in, V_2 will remain cut off because the action of V_1 prevents the plate potential of V_2 from rising to the point of tube ionization.

In the explanation of this circuit it is assumed that the radar transmitter is turned off. There will be minus 30 volts bias applied to V_1 because of the 30-volt drop

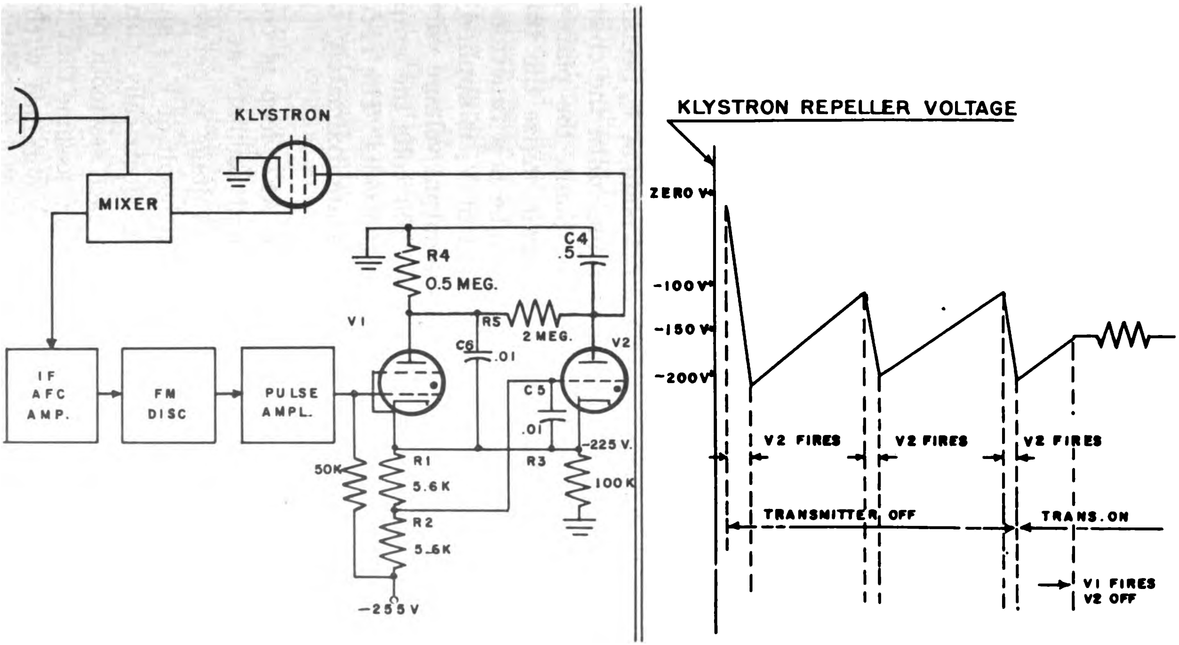


Figure 4–37.—The thyratron AFC control circuit.

across R_1 and R_2. This IR drop is a function of the divider network R_1, R_2, and R_3. V_1 will remain cut off because there will be no signal input to the grid until the transmitter is turned on. Since C_5 cannot be charged instantaneously, the moment that power is applied V_2 will have zero volts bias and a plate-to-cathode potential of plus 225 volts. As a result of these potentials, the gas in V_2 will ionize and the tube will conduct. During the time that V_2 conducts, C_4 will charge towards a minus 225 volts through the low internal resistance of V_2. During this charging period the plate-to-cathode voltage of V_2 will decrease. (The de-ionization potential of a gas tube is a function of the type of gas, the pressure, and the fixed geometrical characteristics of the tube.)

The tube used in this circuit, V_2, will de-ionize at 25 volts. Thus, when the charge on C_4 has reached 200 volts, V_2 will de-ionize. When this occurs, C_4 discharges through R_4 and R_5 to ground and the plate of V_2 becomes less negative. The discharge continues until the charge on C_4 is about 100 volts. At this point, the plate-to-cathode potential is sufficient to again ionize the tube. (The ionization potential of a gas tube is a function of its grid bias.) The sawtooth output of V_2 is applied to the klystron repeller and since this output voltage varies between a minus 100 and a minus 200 volts the amplitude of the sawtooth waveform is 100 volts. The recurrence frequency of this sawtooth is approximately one cycle per second.

To aid in understanding how the AFC loop is completed, assume that the klystron will oscillate at the proper frequency when its repeller voltage is between minus 170 and minus 140 volts. As a result of V_2's sawtooth output, the oscillator will be electrically tuned through the proper frequency during each sawtooth period. When the transmitter is turned on, assume that the proper intermediate frequency will be obtained with a repeller voltage of minus 155. When the repeller is more negative than minus 155 volts the input to the grid of

V_1 will be negative. Thus, V_1 will remain cut off. The discharge of C_4 causes the repeller voltage to become less negative.

As this voltage becomes less negative than minus 155 volts, the input to the grid of V_1 will be positive pulses. At some frequency the pulses will be of sufficient amplitude to ionize V_1, thus providing a new charging path for C_4. This path extends from the minus 225 volts on the thyratron cathodes through V_1 and R_5 to C_4. As C_4 again charges, the repeller voltage will again become more negative; when this voltage passes through the minus 155-volt point, the input to V_1 will again become negative pulses. Thus V_1 will de-ionize as soon as C_6 discharges through it. The de-ionization occurs when C_6 has discharged to the de-ionization potential of V_1 which is about 25 volts. As a result of C_6 maintaining V_1 in conduction for a period of time, it effectively increases the width of the pulses applied to the grid of V_1. After V_1 has de-ionized, C_4 will again discharge through R_4 and R_5 to ground. This entire cycle will be repeated at a rate of about 200 c.p.s.

The amplitude variation of the repeller voltage is so slight that the variation in klystron frequency is only a small fraction of the receiver's IF bandwidth. Because of this action the receiver will remain tuned, or locked-in, to the transmitter frequency. When the circuit is locked-in, V_1 will be conducting periodically and V_2 will remain cut off. Thus V_2 is effectively disconnected from the circuit. The only purpose of V_2 is to provide a slow-continuous sweep to hunt for the proper operating point and to reestablish circuit operation. This action will occur whenever AFC operation is interrupted.

A BEACON AFC CIRCUIT.—When a radar is on beacon operation the receiver must be tuned to some specific frequency whether or not signals are being received. This requires an absolute frequency hunting AFC system for reliable operation. The diagram shown in figure 4–38 is typical of such systems. The output of the klystron is

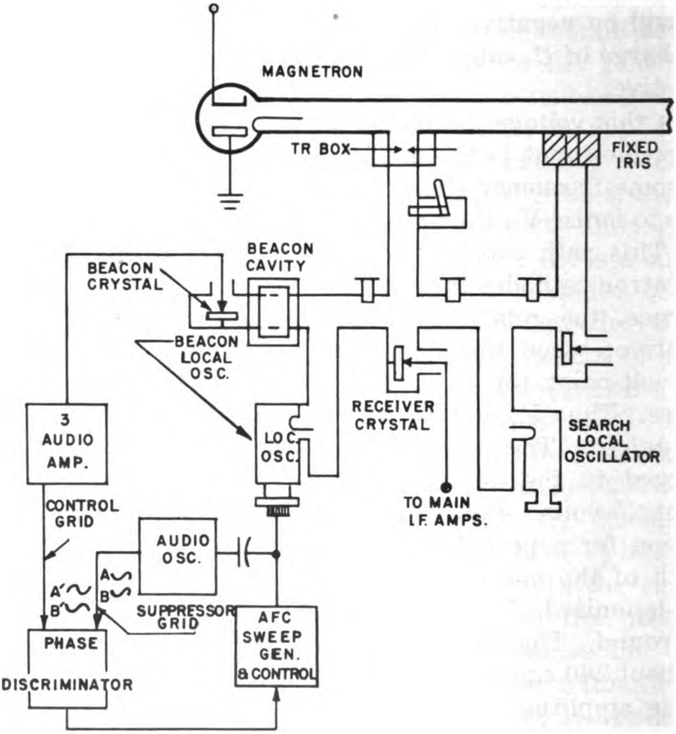


Figure 4–38.—Block diagram beacon AFC loop, using slope discriminator.

coupled through a precision tuned cavity to a separate crystal detector. The filtered output of this detector is a sine-wave voltage (*A′* or *B′*) which is amplified by an audio amplifier and applied to the control grid of a phase discriminator (sometime referred to as a coincidence tube).

The AFC sweep generator and control circuits (see fig. 4–38) used in beacon operation are similar in operation to those used in a radar receiver such as the one just described. Some radar systems use the same AFC sweep generator and control circuits for both radar and beacon. In the following discussion you will note that the phase discriminator, beacon cavity, and other associated circuits replace the frequency discriminator during beacon

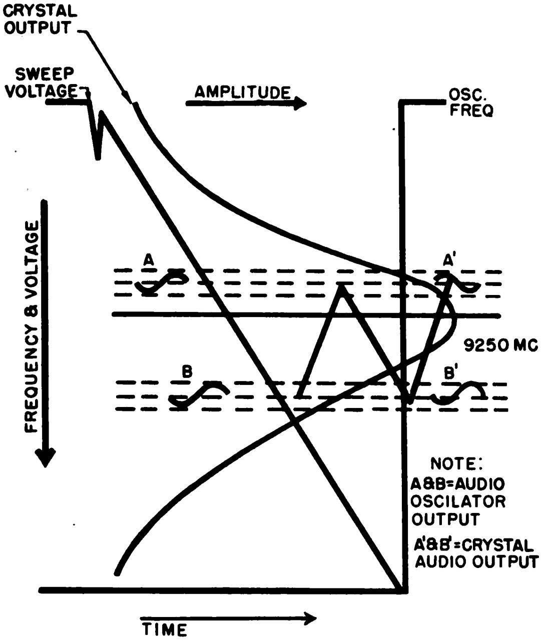


Figure 4–39.—Interrelation of AFC sweep generator and beacon cavity response.

operation. This provides the proper input to the AFC control tube for locked-in operation.

In addition to the sweep control circuit, such as used in a thyratron AFC system, the output of a one-thousand cycle audio oscillator (*A* or *B*) is coupled to the repeller of the klystron. (The amplitude of this audio signal is small, about 0.5 volt.) Thus, the klystron is not only swept over a wide range but it is frequency modulated at an audio rate. The sine-wave output of the crystal detector which is derived by slope detection, will depend upon the frequency response characteristics of the cavity. With reference to figure 4–39, note that when the local oscillator is below the cavity frequency a decrease in local oscillator frequency will result in reduced output from the crystal. An increase in local oscillator frequency will result in an increase in output from the crystal. Thus by SLOPE DETECTION—the process just described—the frequency modulated output of the klystron is converted into an amplitude-modulated signal. The phase of this signal reverses whenever the local oscillator is swept through the resonant frequency of the cavity. Phase shifts to the audio signals are also shown in figure 4–39 at points marked *A*, *A′* and *B*, *B′*.

The resultant variations on the control grid of the phase discriminator are shown in figure 4–38. Notice that when the control grid of the phase discriminator is going positive (*A′*) its suppressor is being driven negative by the audio oscillator output (*A*). Since the potentials on the elements of the phase discriminator tube are so arranged that it will not conduct unless both the control and suppressor grids are driven positive, the tube will remain cut off. The AFC control circuit will continue to increase the klystron frequency. When the local oscillator is above the cavity center frequency a decrease in local oscillator frequency will cause the output of the crystal to increase. When the control grid of the phase discriminator is driven in the positive direction by the audio output (*B′*) the audio oscillator is also driving the suppressor positive (*B*) and the phase discriminator will conduct. The output pulse of the phase discriminator is coupled to the sweep control circuit and will start its locked-in action. The result is a locked-in sweep which holds the klystron near the resonant frequency of the cavity. Notice the similarity between the action of the cavity and the frequency discriminator of the previous AFC radar control circuit.

DELAY SYSTEMS

The term DELAY SYSTEM refers to special control circuits that are used in many electronic systems particularly in radar sets. These circuits are used for the purpose of delaying a waveform, which, in turn, produces a delayed action. A knowledge of several of the basic circuits will prove helpful to you in gaining an understanding of the application and operation of delay systems.

Some of the more common types of electronic delay systems are the R–C, the phantastron, and the delay line (artificial lines).

Purpose and Application

The PPI radar scope is an equipment component that makes use of various applications of different delay systems. Some of these are:

1. Closed center operation. In this operation the PPI sweep is started after the magnetron is fired. (See fig. 4–40.) As compared to normal operation, the targets appear to be moving in toward the center of the scope, hence the name closed center. This is used to eliminate the ground or sea return in the center of the scope. Also, it extends the range of the radar by the amount of delay that is used.

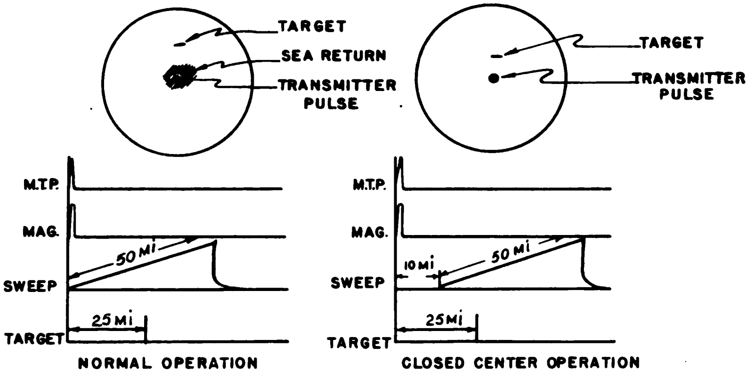


Figure 4–40.—Normal and closed center operation.

2. Open center operation. The magnetron fires after the sweep has started. (See fig. 4–41.) As compared to normal operation, the targets appear to be moving toward the outer edge of the scope. This action moves close-in targets toward the outer edge of the scope where they will occupy more space on the scope face. This spreads the target over more area and will improve target resolution. The effective range on this sweep is reduced by the amount of delay used.

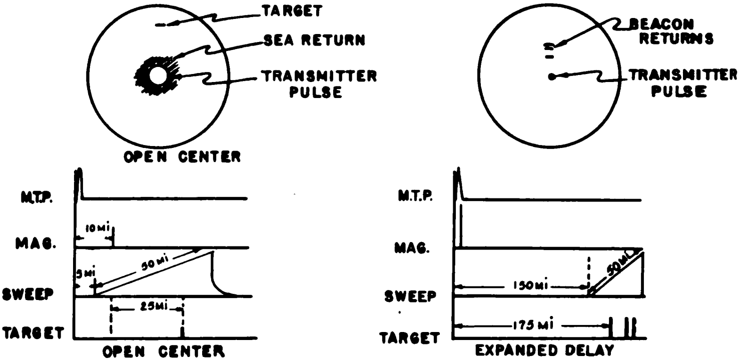


Figure 4–41.—Open center and expanded delay operation.

3. EXPANDED DELAY SWEEP. The start of the sweep is delayed for a number of miles. (See fig. 4–41.) A much faster sweep speed is used but sweep amplitude remains the same. This causes the distant returns to be spread so they can be distinguished from each other. It is also used to aid the operator in reading beacon codes at great distances. The disadvantage is that nothing can be seen on the scope prior to the time the sweep is started. (All close-in targets will be obscured).

4. DELAY LINES. Delay lines may be used for accurate short periods of delay such as delaying the start of a PPI sweep. This enables the sweep and transmitter to be triggered at the same time. They are also used to generate coded pulses. Special types of pulse equipment, in which pulses are transmitted at one location and received at another, frequently use coded pulses. These pulses are used for the purpose of preventing signals of the same carrier frequency, that contain similar pulses, from interfering with the desired signal pulses. A common type of coding is that in which a group of pulses is used for each actual pulse. The pulses in each group are separated by accurate and unequal time intervals. The means most frequently used for generating the pulses is a circuit that contains the delay line (artificial transmission line).

R–C Delay System

This system may be used for open or closed center operation. It is used in some radar systems to permit the sweep to become linear before the transmitter is fired. The PPI sweep starts at time zero, the R–C delay causes the transmitter to be fired later, at a fixed time, after the sweep has become linear.

CIRCUIT ANALYSIS.—(Refer to fig. 4–42.) Since the grid is connected to a positive voltage, V_1 will normally conduct heavily. Thus, plate voltage is low and C_2 will be charged to this low voltage. The input to V_1 is derived from the multivibrator timer and is a negative square wave. The leading edge of this square wave is time zero, and it is at this time that the sweep generator is triggered. The leading edge of this square wave cuts off V_1 which was normally conducting. Plate voltage of V_1 can-

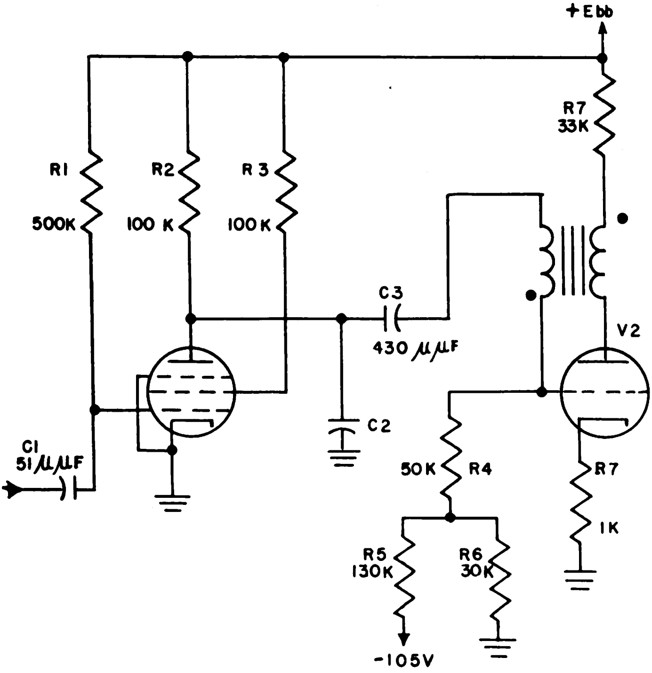


Figure 4–42.—R–C delay circuit.

not rise instantly because of the time required to charge C_2. Although C_2 charges at an R–C rate, during the first 10 percent of time the change in voltage is linear. V_2 is normally cut off by a voltage divider network in the grid circuit which places a minus 19.5 volts on the grid. The linear change in voltage at the plate of V_1 is coupled to the grid of V_2 causing it to go positive and overcome its bias. The delay is a function of the bias applied to the grid of V_2 and the rate of charge of C_2. When the bias is overcome, V_2 conducts and the blocking oscillator action produces a pulse output. A positive pulse output may be taken from the cathode and a negative pulse from across the plate load resistor. The blocking oscillator action in the grid circuit charges both C_2 and C_3 in a negative direction. This causes the plate voltage of V_1 to decrease and keeps V_1 from again triggering V_2 immediately. Before C_2 can again become positively charged to the point that it can trigger V_2, the short R–C in the grid of V_1 has allowed V_1 to conduct. Therefore, the circuit has returned to the static condition of V_1 conducting heavily and V_2 is cut off.

The output of V_2 can be used to start the sweep and other timing circuits such as range-mark generators.

Phantastron Delay System

A phantastron delay system may be used to produce a variable delay over the PPI range. In order to provide a delay over most of the full range of the PPI, it is necessary to use a cathode follower with the phantastron.

CIRCUIT ANALYSIS.—The phantastron part of the circuit in figure 4–43 is similar to the one described earlier in this chapter and shown in figure 4–25. The phantastron is triggered simultaneously with the firing of the transmitter. The trigger input is applied to the cathode of *V*118 and the output of the phantastron is taken from the screen of *V*119. This output is a square wave that goes positive at the start of the delay and negative at the end. The output of the phantastron is applied to the trigger amplifier *V*121*A*. The end of the delay ap-

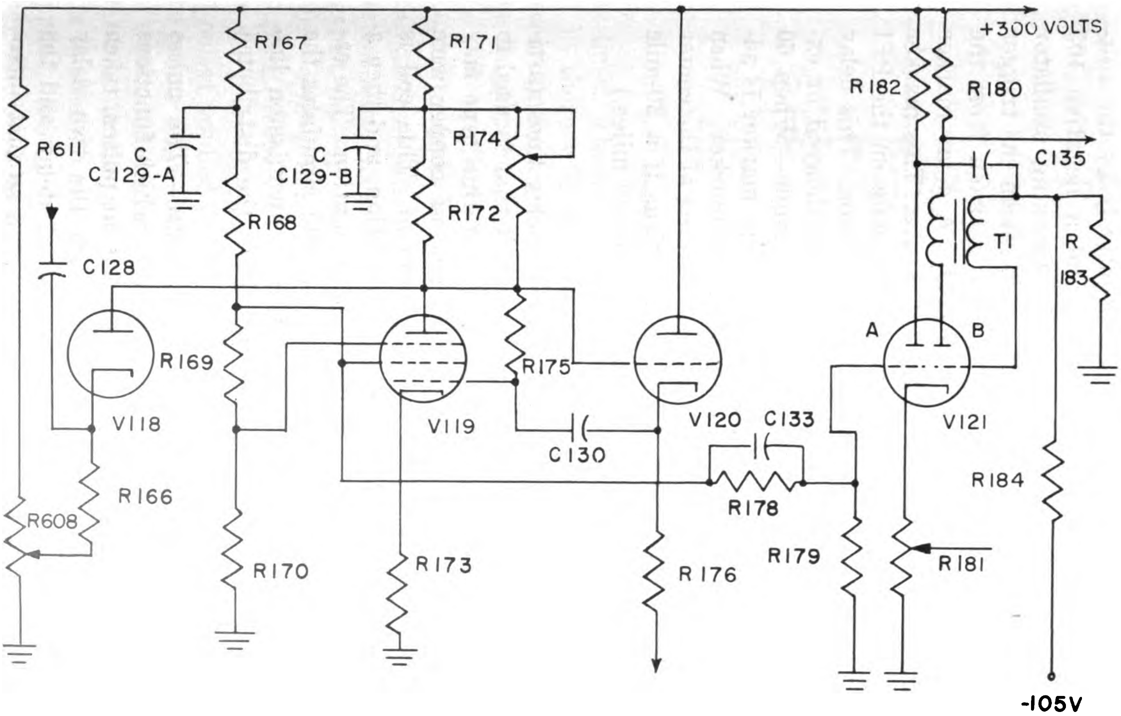


Figure 4—43.—Phantastron delay system.

pears as a positive pulse at the plate of the amplifier. This pulse is coupled to the grid of the blocking oscillator *V*121*B*. This oscillator is biased negatively by the voltage divider network *R*183, *R*184, and a negative 105 volts; thus, it is normally cut off. Blocking oscillator action is initiated by the positive pulse from the trigger amplifier. The delayed sweep trigger is taken from the plate of the blocking oscillator *V*121*B*. A delay marker is taken from the cathode of the same tube. This marker appears as a movable range marker or ring on the PPI when the scope is in a nondelayed condition. This delay marker indicates the point at which the delayed or expanded sweep will start. (NOTE: Example—When 60 miles of sweep delay is desired, the delay marker is adjusted to aline with the 60-mile range marker. When the delay is switched-in, the sweep will start at the center of the scope after 60 miles of delay. Thus if a 30-mile sweep is used, the scope will cover 60 to 90 miles.)

Delay Line System

Figure 4–44 shows an application of delay lines (artificial transmission line type) in a circuit that is used to generate coded pulses. A typical line of this type may be constructed by winding many turns of copper wire on a flexible tube of insulating material. This coil is first covered with a thin layer of insulation, and then a braided shield and another layer of insulation. The capacitance between the coil and the shield simulates the distributed capacitance of the actual transmission line and the inductance of the coil simulates the distributed inductance.

CIRCUIT ANALYSIS OF A CODING SYSTEM.—The pulses to be coded are coupled to the grid of V_1 which functions as a blocking oscillator. The sharp output pulses, taken from the cathode of V_1, pass through the two delay lines. One of these lines has 2 μsec. of delay and the other 4 μsec. Each delay line is designed so that introduction of a single pulse will result in an output of two pulses spaced according to the characteristics of the de-

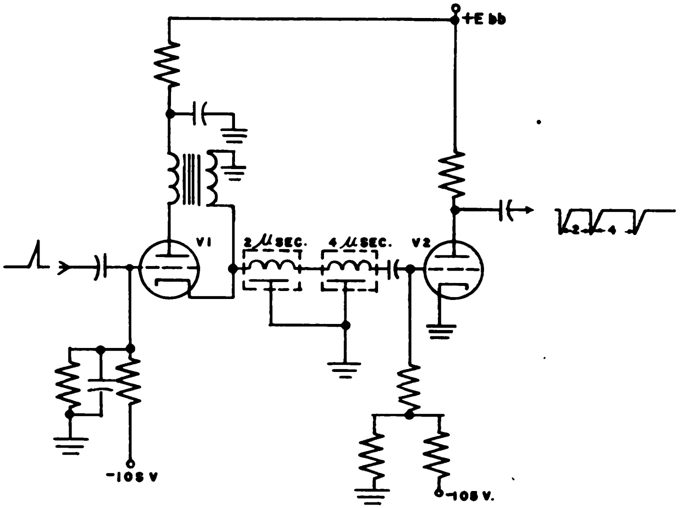


Figure 4–44.—Delay line coding circuit.

lay lines. Thus, for each pulse output from the cathode of V_1 there will be three pulses on the grid of V_2. The second pulse will occur 2 μsec. after the first pulse and the third pulse will occur 6 μsec. after the first pulse. V_2 will amplify and invert the pulses, they are then coupled to the associated circuits.

CIRCUIT ANALYSIS OF A DECODING SYSTEM.—A decoding system must be employed at the receiver to discriminate against undesirable pulses. A train of properly spaced pulses will provide the coincidence necessary for these to pass through the decoding circuits. Thus, there will be one pulse at the output of the decoder for each train of pulses applied to the input. Pulses that do not possess the proper time spacing will be eliminated by the decoder. A typical decoding circuit is shown in figure 4–45. In this circuit the pulses taken from the secondary of T_1 are applied to the three grids of V_1. These pulses are applied directly to grid 3, but because of the delay lines they will be delayed 4 μsec. to grid 2 and 6 μsec. to grid 1. V_1 is so biased that the application of any one or even two pulses, to the grids will not cause the tube to con-

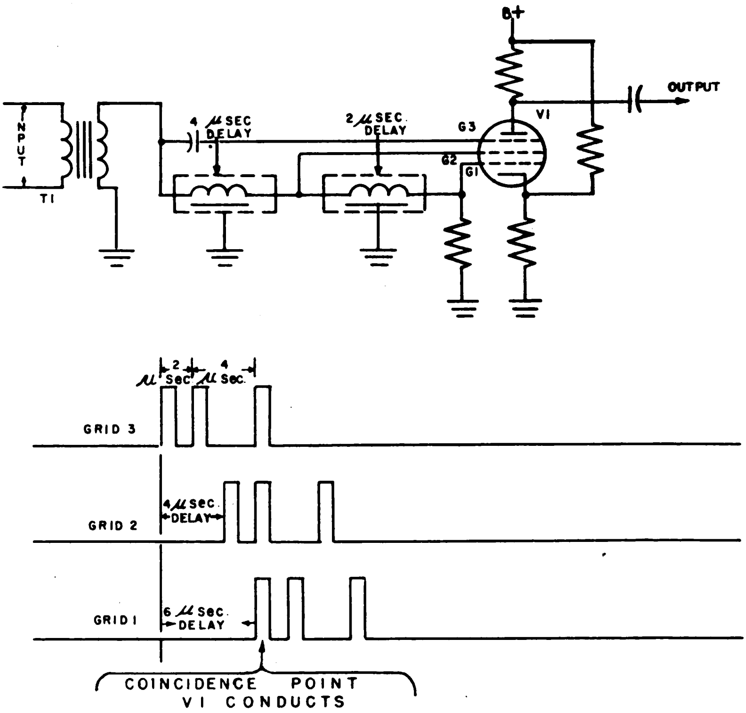


Figure 4–45.—Delay line decoder.

duct. When three pulses are applied simultaneously to the three grids the bias is overcome and the tube conducts. This coincidence is graphically illustrated in figure 4–45. By referring to this graph you can easily see, as a result of the pulse spacing used, that at no time other than the coincidence will more than one pulse be applied to V_1.

MARKER SYSTEMS

Marker systems are used to enable the radar operator to determine the location of targets with great accuracy. This may be accomplished by causing visual guides to appear on the face of the scope. These guides or markers will provide range or distance to a target and azimuth bearing of a target. Some systems present additional information that gives the operator the target height.

Markers are also used on equipment other than radar, for example—electronic countermeasures, loran, oscilloscopes, and many others. In these equipments, markers are used to determine pulse width, pulse spacing, pulse amplitude, and various other functions.

Early radars depended on the linearity of the scope sweep for accuracy in range. A rough range scale was etched on a transparent plastic plate and this was placed on the face of the scope. Present-day radars obtain range marks from a range-mark circuit. This circuit consists of a stable oscillator that generates accurately spaced range marks which are used to modulate the electron beam in a PPI type scope. This type presentation causes intensity type markers to appear.

The type A scope may also utilize intensity type marks or else the markers may be applied to the deflection circuits. The application of the marker to the deflection circuits causes pulse type markers to appear on the scope.

Fixed Range Marker

This circuit uses a gated type ringing oscillator which produces range marks that are synchronized with the sweep. A circuit that illustrates this system is shown in figure 4–46.

CIRCUIT ANALYSIS.—The gating tube V_1, because of positive bias, is normally conducting and producing a steady flow of current through L_1. The flow of current through L_1 maintains a magnetic field around the coil; thus, the tank circuit composed of L_1 and C_2 cannot oscillate. Simultaneously with the starting of the scope sweep a negative square wave is applied to the grid of V_1, cutting the tube off and causing the magnetic field around L_1 to collapse. The collapsing of this field shock-excites the oscillator. The circuit will continue to oscillate throughout the duration of the negative square wave. The values of L and C, in the series-fed Hartley oscillator, are so chosen that the duration of each cycle of oscillation will be equal to the time between the desired range mark spaces. The leading edge of each negative

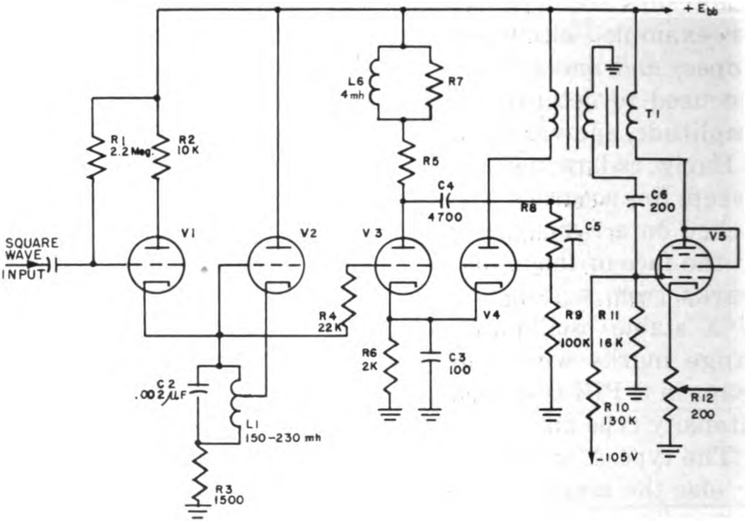


Figure 4–46.—Ringing oscillator range marker.

oscillation will produce a range mark in the output of the circuit. V_2 replenishes the losses of the tank circuit thereby maintaining the amplitude of oscillation.

The peaking amplifier V_3, that is normally conducting, is cut off by the oscillator's negative output. A high positive voltage at the plate of V_3 will be produced by the collapsing magnetic field around L_6. Oscillations across L_6 are prevented by damping resistor R_7, thereby preventing possible double range marks on the scope. Positive oscillation of the grid signal causes limiting to occur in the grid of V_3 due to R_4 in the grid circuit. Therefore, the negative-going portion is the only part of the input wave that has an appreciable effect on the plate current of V_3. The positive pulse output is coupled to the grid of V_4 through C_4 and C_5. C_4 functions as an ordinary coupling capacitor; C_5 in conjunction with R_8 acts as a fast coupling network. The leading edge of the output pulse from V_3 contains high frequency components; these are coupled through C_5 which bypasses R_8, therefore the complete leading edge of the output of V_3 is developed across R_9. This sharp rise in voltage on the grid of V_4 causes it to draw current charging the lower plate of C_5 negatively, since C_5 and R_8 have a R–C time of approximately 1 second. The charge due to grid current rapidly discharges through R_8 and returns the coupling circuit to a static condition.

If only C_4 were used to couple the sharp rise to the grid of V_4 when the grid drew current, the grid side of C_4 would go negative. Since the discharge path for C_4 is long, it would soon assume an average charge and the coupled signal would be reduced to minimum. The blocking oscillator V_5 is normally cut off by a fixed negative voltage on the grid. This circuit is triggered by the output of V_4 which is coupled to the grid of V_5 by transformer T_1. Positive range marks are taken from the cathode of V_5.

Movable Range Marker

Many present day radarscopes utilize the movable marker or RANGE STROBE. This system provides a continuously variable marker that may be used to aid the operator in quickly determining the accurate range of a target. Movable marker action may be incorporated in a radar system by using mutivibrators or the phantastron time delay circuits previously discussed in this chapter. Another system whereby such marker action may be obtained is a timing circuit using a linear sawtooth and diode pickoff tube (fig. 4–47). This system is explained in detail in the following paragraphs.

CIRCUIT ANALYSIS.—The sweep generator tube V_1, which is normally conducting, is cut off by the negative input square wave. When cut off, its plate voltage will rise at a rate determined by the sweep capacitor C_1. In order to obtain a linear sawtooth output it is necessary to use either the first ten percent of the charge or to add a bootstrap circuit to the sawtooth generator. This sawtooth waveform is applied to the plate of V_2. V_2 is normally cut off by the positive voltage on its cathode. The sawtooth rise applied to the plate of V_2 eventually overcomes the cathode voltage and causes the diode to con-

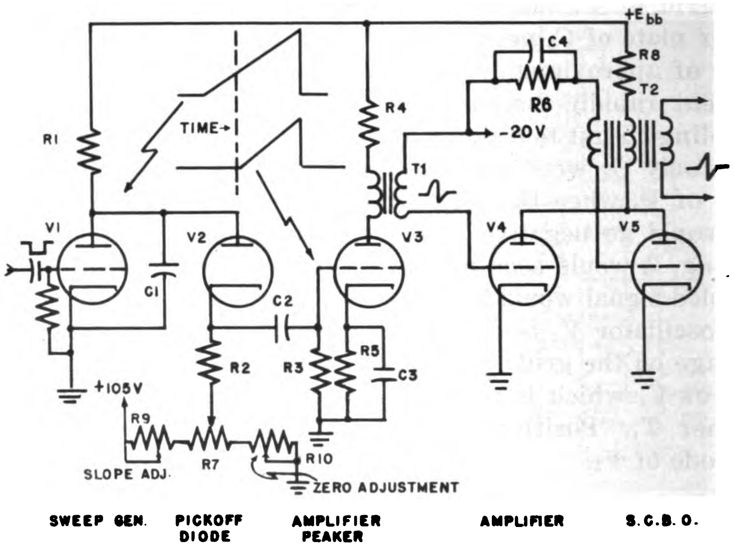


Figure 4—47.—Sawtooth type movable marker.

duct. When this occurs the cathode voltage rises at the same rate as the plate voltage and is applied to the grid of V_3. Tube V_3, because of the transformer T_1 in its plate circuit, then functions as a peaker amplifier. The output of V_3 is applied to the grid of V_4 and causes it to conduct momentarily. The current through the plate winding of T_2, due to the conduction of V_4, causes the single-cycle blocking oscillator to cycle and produce the delayed pulse output.

The duration of the delay depends upon the voltage applied to the cathode of V_2; this voltage is determined by the position of the potentiometer arm of R_7. As the arm of the potentiometer is moved toward the more positive end, a higher voltage is then applied to the cathode of V_2 and more time is required for V_2 to start conducting, thus giving more delay before the pulse is produced. This delayed pulse is the movable marker that appears on the scope. Range is read from a mechanical counter or other indicating device which is mechanically coupled to the arm of the delay potentiometer R_7. This counter is calibrated by adjusting the zero adjustment R_{10} and the slope adjustment R_9. These potentiometers are positioned to provide the correct voltage across R_7 so that the delay that is provided, dependent upon the location of its arm, is correctly indicated by the counter.

Movable Height Marker

The linear sawtooth delay circuit (fig. 4–47) may be used as the timing system for measuring target height with radar equipment. This circuit was explained in detail under the heading "Movable Range Marker" in this chapter. In order for it to apply to movable height markers certain modifications must be made. The angle of elevation of the height finder antenna is referred to as Φ. In the movable range marker (fig. 4–47), a fixed supply potential was used for the sawtooth generator, but when this circuit is used for height finding the supply voltage varies with the angle of elevation of the antenna. Thus, the slope of the sawtooth is proportional to sin Φ. In this application the counter or other indicator that is connected to the delay potentiometer indicates the target's height.

The curvature of the earth may be neglected when determining height at short or medium ranges. The relationship between delay time and target height, or the time when the height marker appears on the sweep, may be expressed as follows:

$$T = \frac{2H}{C \sin \Phi}$$

T equals time

H equals height

C equals speed of radio waves

Φ equals angle of elevation

Since the slope of the sawtooth has been made proportional to sin Φ, the time delay will be inversely proportional to sin Φ or, in other words, directly proportional to $\frac{1}{\sin \Phi}$ for a given diode bias.

QUIZ

1. A multivibrator employs tubes as
   a. electronic switches
   b. amplifiers
   c. detectors
   d. oscillators

2. In a multivibrator circuit the frequency-determining factors are
   a. the types of tubes employed
   b. the tank capacitors
   c. grid capacitors and resistors
   d. the values of the plate loading resistors

3. In figure 4–1, increasing the size of C_2 will
   a. increase output frequency
   b. increase time of negative portion of waveform on the plate of V_2
   c. decrease frequency
   d. increase amplitude of waveform on plate of V_2

4. In figure 4–1, increasing R_3 will
   a. increase frequency
   b. increase amplitude of waveform at plate of V_1
   c. increase time of positive portion of waveform at plate of V_1
   d. decrease amplitude of waveform at plate of V_1

5. In a free-running multivibrator (see fig. 4–1), returning the grids to a positive potential will
   a. decrease frequency
   b. improve frequency stability
   c. increase amplitude of waveform at plate of V_2
   d. decrease amplitude of waveform at plate of V_1

6. An operating free-running multivibrator differs from an operating one-shot multivibrator in having
   a. feedback between stages
   b. two stages, alternately conducting and nonconducting
   c. a synchronizing waveform required for operation
   d. no equilibrium condition

7. The cathode-coupled one-shot multivibrator is like the plate-coupled multivibrator in that
   a. each stage is coupled regeneratively to the other stage in order to produce rapid action
   b. feedback action is from the plate of each stage to the grid of the other stage
   c. two stages return to their initial states of conduction automatically in order to be ready for the next trigger
   d. feedback action of the second stage results from the cathode-to-cathode connection

8. In figure 4–17, increasing R_2 will
   a. increase time of negative portion of output
   b. increase amplitude of output
   c. decrease amplitude of output
   d. increase time of positive portion of output

9. A multivibrator can be synchronized by applying
   a. only positive pulses to the conducting tube
   b. only negative pulses to the nonconducting tube
   c. either positive pulses to the nonconducting tube or negative pulses to the conducting tube
   d. either positive pulses to the conducting tube or negative pulses to the nonconducting tube

10. Frequency division, as obtained by a multivibrator, refers to the operation of a driven multivibrator which is
   a. driven at a submultiple of its free-running frequency
   b. driven at a harmonic of its free-running frequency
   c. synchronized to a harmonic of the driving frequency
   d. synchronized to a submultiple of the driving frequency

11. In figure 4–17, output frequency could be reduced to one-half the input by
   a. decreasing the size of C_2
   b. increasing the size of C_2
   c. decreasing the size of R_2
   d. increasing the RC time of R_1C_1

12. What characteristic of the input pulse to a positive step-by-step counting circuit can vary while the circuit is in operation without affecting the calibration of the equipment?
    a. Shape
    b. Amplitude
    c. Repetition rate
    d. All of the above

13. In a scale-of-two multivibrator, with positive pulses applied to both grids, the output frequency will be
    a. twice input frequency
    b. equal to input frequency
    c. one-half input frequency
    d. always higher than input frequency

14. In a phantastron delay circuit, the pulse width is normally increased by
    a. decreasing plate voltage
    b. increasing plate voltage
    c. increasing the value of the plate load resistor
    d. decreasing the value of the plate load resistor

15. A cathode follower may be added to a phantastron circuit to
    a. provide a more linear delay
    b. provide short delay periods
    c. prevent the following circuit from loading the phantastron circuit
    d. provide long delay periods

16. In a blocking oscillator, the tube is said to be blocked when the
    a. plate current reaches saturation
    b. grid bias has reached its most positive value
    c. grid is beyond cutoff and no plate current flows
    d. grid is beyond cutoff and maximum plate current flows

17. In a free-running single-swing blocking oscillator, the frequency is controlled by the
    a. resonant frequency of the transformer
    b. *RC* time in the grid circuit
    c. type of tube used
    d. size of the cathode resistor, when used

18. In figure 4–31, increasing input frequency will cause
    a. voltage across R_1 to increase
    b. voltage across R_1 to decrease
    c. meter "M" to decrease its reading
    d. charge on C_3 to decrease

19. In figure 4–34, to increase the count-down ratio,
   a. decrease bias on V_2
   b. decrease value of C_1
   c. increase value of C_1
   d. decrease value of C_2

20. In figure 4–37, when the klystron is operating at the proper frequency,
   a. V_1 will conduct periodically
   b. V_2 will conduct periodically
   c. the charge on C_6 will remain constant
   d. the charge on C_4 will remain constant

21. The type of PPI presentation used to read beacon codes of great distance is
   a. open center operation
   b. expanded delay sweep
   c. normal sweep (no delay)
   d. closed center operation

22. The oscillatory circuit, which generates the range markers in most radar sets, is gated by a
   a. sawtooth pulse
   b. sine wave
   c. trapezoidal wave
   d. square wave

CHAPTER 5

WAVEGUIDES AND CAVITY RESONATORS

The frequencies employed by microwave equipments, such as radar, make possible the use of two unique, but very practical devices—waveguides and cavity resonators. A WAVEGUIDE is a hollow pipe for transferring high-frequency energy. A CAVITY RESONATOR is a hollow metallic cavity in which electromagnetic oscillation can exist when the cavity is properly excited. The purpose of this chapter is to acquaint you with these devices. It discusses their theory, operation, and uses.

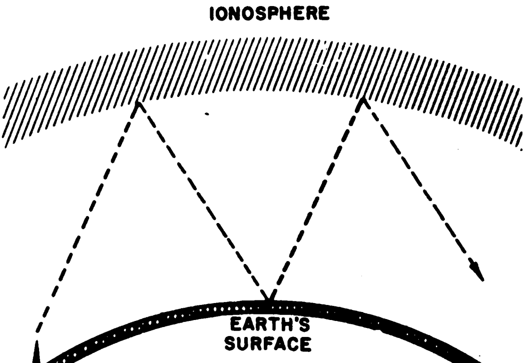
WAVEGUIDES

Guided and Unguided Electromagnetic Waves

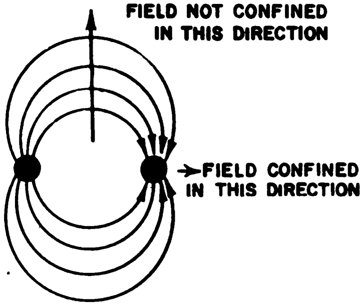
In general, there are two methods for transferring electrical energy: one is by current flow in conductors; the other is by movement of electromagnetic fields in space. Electrical energy can be transferred as current flow in a number of types of transmission lines, for example, two-wire lines and coaxial lines. In space, energy is transferred by moving electromagnetic fields, an example being the radiation of energy from a radio antenna. Many electromagnetic fields which move in space are confined largely to the area between the earth and the ionosphere, as shown in (A) of figure 5–1. (The ionosphere is the thick layer or cloud of free ions and electrons which exists at a height of about 60 miles above the earth.) The change in the dielectric constant of this part of the atmosphere is sufficient to refract electromagnetic fields which strike it. If this refraction is great enough, the fields will be returned to the earth. All electromagnetic fields are not returned, for extremely high-frequency radiations that strike it almost perpendicularly penetrate the ionosphere and are lost. An electromagnetic field that does not start out in a direction parallel to the earth and the ionosphere follows a zigzag path between these two areas and may or may not be reflected back to the earth.

Although the transfer of energy by electromagnetic fields and by currents in conductors may seem to be unrelated phenomena, actually at present, electronics scientists tend to regard even two-wire lines as elements which guide electromagnetic fields from one place to another. The currents in the wires are considered incidental to the action and the result of the moving fields.

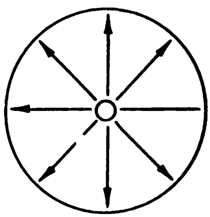
A two-wire line is a poor guide for transferring electromagnetic fields, because it does not confine the fields in a direction perpendicular to the plane which contains the wires. (See (B) of fig. 5–1.) This results in some energy escaping in the form of radiation. Electromagnetic fields may be completely confined in this direction when one conductor is extended around the other to form a coaxial cable, as shown in (C). In a coaxial cable, energy transfer also is said to take place by the motion of electromagnetic fields, rather than by current flow. However, this method is not very efficient at high radiofrequencies since skin effect limits the current carrying area of a conductor to a thin layer of its surface. Another disadvantage is that the ability of the field to form is limited by the amount of current flow associated with it. When the resistance in a conductor is increased, current flow is reduced accordingly. This reduces the magnitude of the fields. On inspecting the cross section of the coaxial cable, you can see that the surface area of the inner conductor is much less than the surface area of the outer conductor. This causes the inner conductor to retard the current considerably more than the outer conductor and results in a reduction in the efficiency of energy transfer. If you could remove the center con-



(A) RADIO ENERGY IS CONFINED BY EARTH AND IONOSPHERE


FIELD NOT CONFINED IN THIS DIRECTION
FIELD CONFINED IN THIS DIRECTION

(B) END VIEW OF TWO WIRE LINE FIELDS CONFINED IN TWO DIRECTIONS BUT NOT IN OTHER TWO



(C) END VIEW OF COAXIAL CABLE FIELDS CONFINED IN ALL DIRECTIONS

Figure 5–1.—Guiding waves.

ductor and retain the fields, energy might be transferred with less loss.

Electromagnetic fields can transfer energy in a line which does not have a center conductor provided the configuration of the fields is changed to compensate for the missing conductor. The object remaining, which is virtually a hollow pipe, is called a WAVEGUIDE. A waveguide does not necessarily have to be circular in cross section. Practical waveguides, for example, are sometimes square, rectangular, or elliptical in cross section.

Metallic walls are not necessary to guide electromagnetic fields in a waveguide, for the fields will be reflected whenever they encounter any kind of a substance which has a different dielectric constant than the substance in which they are traveling. For example, fields can be made to travel through a ceramic rod with only a small loss of energy. When they encounter the air at the surface of the rod, they are reflected back into the rod.

Waveguides Vs. RF Lines

As you previously learned, the three types of losses in RF lines are copper losses, dielectric losses, and radiation losses. Briefly these losses are described as follows: Copper loss is an I^2R loss. It becomes appreciable whenever skin effect reduces the conducting area of the lines. Dielectric losses, as you recall, are losses due to the heating of the insulation between conductors. Radiation losses are losses due to energy escaping from RF lines in the form of radiation.

ADVANTAGES OF WAVEGUIDES.—Considering waveguides from the point of view of these losses, they have the following advantages:

1. Copper losses are small in waveguides. Since a two-wire line consists of a pair of conductors which are small, the surface area of each is likewise small. Although the surface area of the outer conductor of a coaxial cable is large, the inner conductor is small and pro-

duces considerable copper losses. On the other hand, a waveguide does not depend on conduction through a center conductor. Therefore, whenever current flows, this flow is in the walls, and copper losses are less than those in other types of lines.

2. Dielectric losses are small in waveguides. In two conductor lines, some form of insulation is used between the conductors. The fields which move around this insulator cause heat, and the heat in turn takes power from the line. A waveguide has no center conductor to support. Furthermore, there is only air in the hollow pipes. Since the dielectric loss of air is negligible, it follows that the dielectric losses in a waveguide are small.

3. Radiation losses are less in a waveguide than in a two-wire line. In a waveguide, fields are contained wholly within the guide itself, just as in a coaxial line. Therefore, only a negligible amount of energy is radiated.

4. The power-handling capacity of a waveguide is greater than that of a coaxial line having an equal size. Power is a function of E^2/Z_0, where E is the maximum voltage in the traveling wave and Z_0 is the characteristic impedance of the line. E is limited by the distance between the conductors. In the coaxial line (fig. 5–2), this distance is S_1. In the waveguide, the distance which is S_2 is much greater than S_1. Therefore, the waveguide is able to handle greater power before the voltage exceeds the breakdown potential of the insulation.

5. A waveguide is simpler to construct than a coaxial line, since the center conductor is eliminated completely.

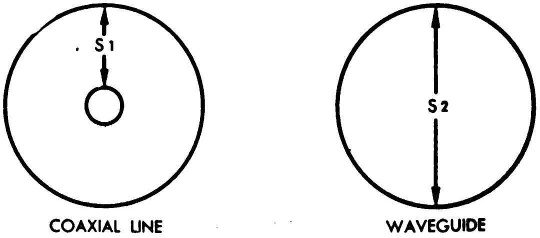


Figure 5–2.—Comparison of spacing in coaxial line and waveguide.

6. The physical stamina of a waveguide is greater than that of a coaxial line. Unlike a coaxial line, it has no center conductor or insulators which can be displaced or broken.

DISADVANTAGES OF WAVEGUIDES.—In view of the advantages, you may wonder why waveguides are not used exclusively for transferring energy. There is, however, one major disadvantage which makes waveguides impractical at any but extremely high frequencies. The cross-sectional dimensions of a waveguide must be in the order of a half wavelength for it to contain electromagnetic fields properly. A waveguide used at one megacycle, for example, would be about 700 feet wide. At 200 mc. this waveguide would have to be about 4 feet wide, while at higher frequencies, such as 10,000 mc., it need be only one inch wide. Therefore, dimensions which waveguides require make them impractical at any frequency lower than about 3,000 mc. If the critical dimension of the guide is less than a half wavelength, energy will not be propagated through the waveguide. The reason for this is that for any given waveguide there is a cutoff frequency, below which it does not function as a power transfer device. This also limits the frequency range of any system using waveguides. Coaxial cables do not have this limitation for they have no cutoff frequency and will transmit low as well as high frequencies.

The installation and operation of waveguide transmission systems are somewhat more difficult than for other types of line. The radius of bends in the line must be greater than two wavelengths to avoid excessive attenuation. This fact may hamper installations in restricted spaces. If the guide is dented, or if solder is permitted to run inside the pipe when joints are made, the attenuation of the lines is greatly increased. In addition to the increased attenuation that they cause, dents or beads of solder also reduce the breakdown voltage of the waveguide. Although such faults may not cause actual arcover in the guide, they limit the power-handling capacity of the system and make the possibility of arc-over more likely. Thus, unless great care is exercised in the installation, one or two carelessly made joints may completely nullify the initial advantage obtained from the use of the waveguide.

WAVEGUIDE THEORY

An exact mathematical analysis of the way in which fields exist in a waveguide is quite complicated. However, it is possible to obtain an understanding of many of the properties of waveguide propagation by using the following simple analogy, which shows how the fields are able to exist in a waveguide and how they can be handled.

Analogy of Waveguide Action From a Two-Wire RF Line

To understand the action of a waveguide, assume that it has the form of a two-wire line. In this condition there must be some means of supporting the two wires. Furthermore, the support must be a nonconductor, so that no power will be lost by leakage resistance. An efficient way for insulating and supporting the two-wire line is shown in figure 5–3 (A). This line is spaced, insulated, and supported by porcelain standoff insulators. At radiofrequencies, the absorption of power by the dielectric material (insulators) causes them to appear as a resistance and capacitance. The equivalent electrical circuit at radiofrequencies is shown in (B) of the figure. For frequencies of 3,000 mc. and above a better insulator than nonconducting porcelain must be used. A superior high-frequency insulator for this purpose is a quarter-wave section of RF line called a metallic insulator. (See fig. 5–3 (C).) Since there are no dielectric losses in a quarter-wave section of an RF line, the impedance at the open end (junction of the two-wire line) is very high.

A metallic insulator can be placed anywhere along a two-wire line. Several insulators on each side of a two-wire line are shown in figure 5–4 (A). A point to note in this line is that the supports are a quarter wave at

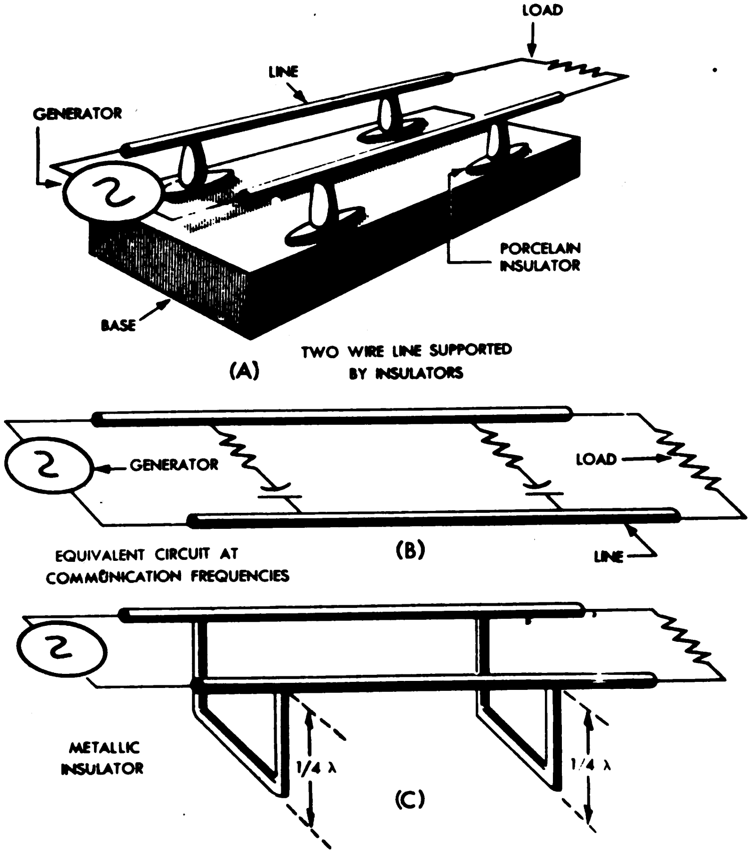


Figure 5–3.—Insulating the two-wire line.

only one frequency. This limits the high efficiency of the two-wire line to only one frequency.

The use of several insulators results in improved conductivity of a two-wire line when the sections are connected together. This connection is made between the two adjacent insulators through a switch, as illustrated in figure 5–4 (B). When the switch is open, both quarter-wave sections are excited by the main line. In this condition, there will be standing waves on the quarter-wave sections. When the switch is connected to

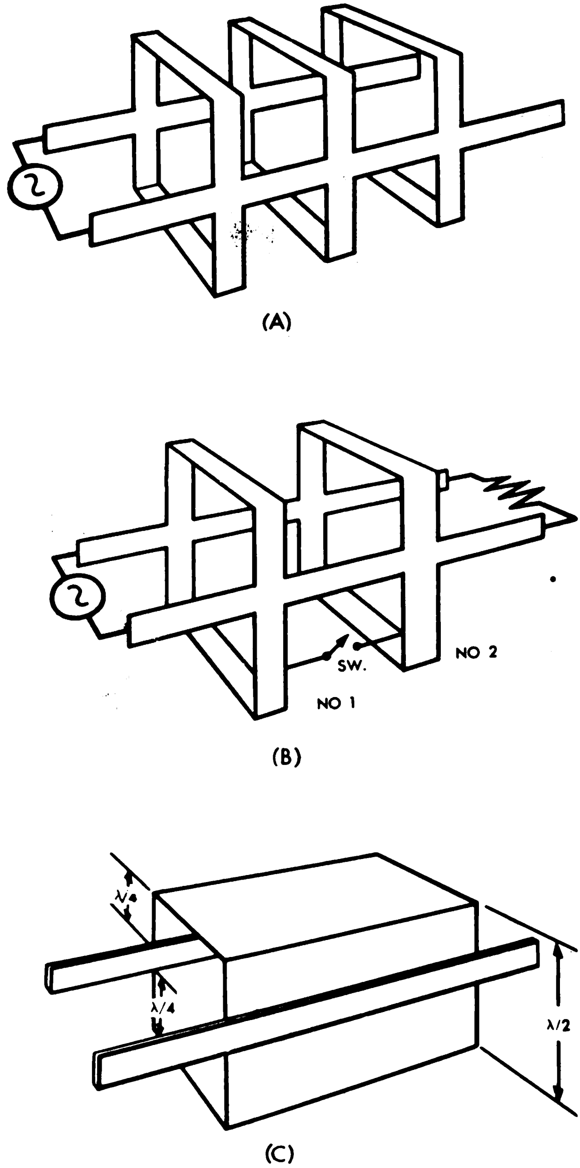


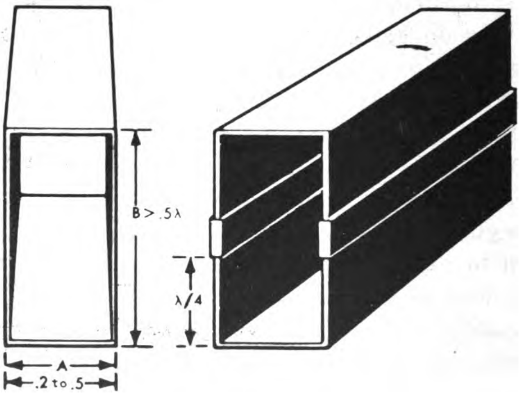
Figure 5–4.—Development of waveguide by adding quarter-wave sections.

the same place on each section, the relative phase relationship of the voltages at the connection will be the same for each section. In this condition, the No. 1 section will be excited first by the generator. When the switch is closed, the No. 2 section will be partly excited by the No. 1 section through the switch connection. In this condition, less energy from the main line will be required to excite the No. 2 section. The parallel paths shown cause less resistance to exist along a given length of line, and energy is transferred with less copper loss.

When additional sections are added to the line until each section makes contact with the next, the result is a rectangular box in which the line is at the center, as shown in figure 5–4 (C). The line itself is actually part of the wall of the box. The rectangular box thus formed is a waveguide.

EFFECT OF DIFFERENT FREQUENCIES ON A WAVEGUIDE.—Previously it was stated that a quarter-wave section is limited in operation to a certain frequency. However, when a solid wall of insulators is added, the section will operate at other frequencies. The waveguide shown in figure 5–5 (A) is the one just discussed. When the frequency being transferred by the guide is made higher, the quarter-wave section must be shorter. These shorter wavelengths are easily accommodated if you assume that this two-wire line is made up of a wide bar or strip in each wall of the guide, as shown in (B). The shorter distance remaining is the shorter quarter-wave section. Thus, the wide bar shown is theoretically well insulated at any frequency higher than the near-minimum frequency illustrated in (A). In reality, there is a practical upper frequency limit to which this analogy is applicable. For example, when the bar is a half wavelength across, the waveguide will be four quarter wavelengths across and may act as though there were two bars instead of one.

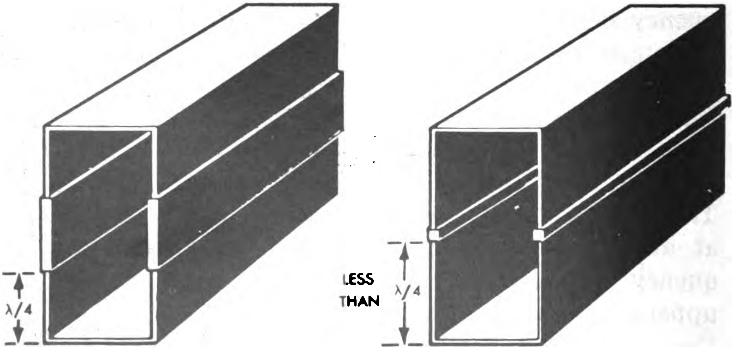
The next consideration is a frequency which is lower than the original frequency. Lowering the frequency in a given waveguide will lengthen the sections and narrow the bar. Below a certain frequency this bar does not exist because the quarter-wave sections meet one another. At a still lower frequency, the sections become less than a quarter wave. (See fig. 5–5 (C).) A shorted section



WAVEGUIDE DIMENSIONS

NEAR MINIMUM FREQUENCY

(A)



ABOVE MINIMUM FREQUENCY

(B)

BELOW MINIMUM FREQUENCY

(C)

Figure 5--5.—Effect of different frequencies in the waveguide.

less than a quarter wave in inductive. Thus, the impedance across the place where the conducting bars belong is NOT a high resistance but an inductance. The inductive reactance will dissipate the energy swiftly through high currents which flow back and forth in the line as though the inductance were taking energy during one half cycle and then returning it to the line during the next half cycle. Thus it follows that in a waveguide there is a low-frequency limit or cutoff frequency, below which the waveguide cannot transfer energy. The width of the waveguide at the cutoff frequency is equal to one half wavelength, since at this frequency the two quarter-wave sections touch and add to one another. Mathematically, the cutoff wavelength is expressed by the equation,

$$\lambda_c = 2B$$

where λ_c is the cutoff wavelength, and B is the long dimension of the rectangular cross section.

Cutoff occurs when the width of a waveguide is below a half wavelength. Most waveguides are then made 0.7 wavelength in the wide dimension to give a margin between the actual size and the size for cutoff. The other dimension is the distance between conductors, and similarly, as in the two-wire lines, is governed by the voltage breakdown potential of the dielectric which is usually air. Widths of 0.2 to 0.5 wavelengths are common.

Electromagnetic Fields in a Waveguide

An understanding of the fields present in a waveguide is necessary in order for the guides to be used effectively. Energy in a waveguide is transferred by the electromagnetic fields, while currents and voltages merely aid in forming these fields. It is important to know when a good current path is required and where the voltage will be high. Also it is important to know where the fields will exist since it is by means of these fields that energy is normally introduced into or removed from the guides.

The operation of a waveguide involves two kinds of fields —the electrostatic and the electromagnetic.

THE ELECTRIC FIELD.—The existence of an electrostatic (electric field) indicates that there is a difference in the number of electrons between two points. The simplest form of electrostatic field is the one which occurs between the two parallel plates of a capacitor. (See fig. 5–6 (A).) When the top plate of the capacitor is made positive by a battery, electrons move from the top plate and deposit themselves on the bottom plate. This immediately sets up a stress in the dielectric between the plates. This stress is represented by arrows, whose direction is from the more positive voltage point to the less positive, or negative, voltage point. The amount of stress is indicated by either the length or the number of

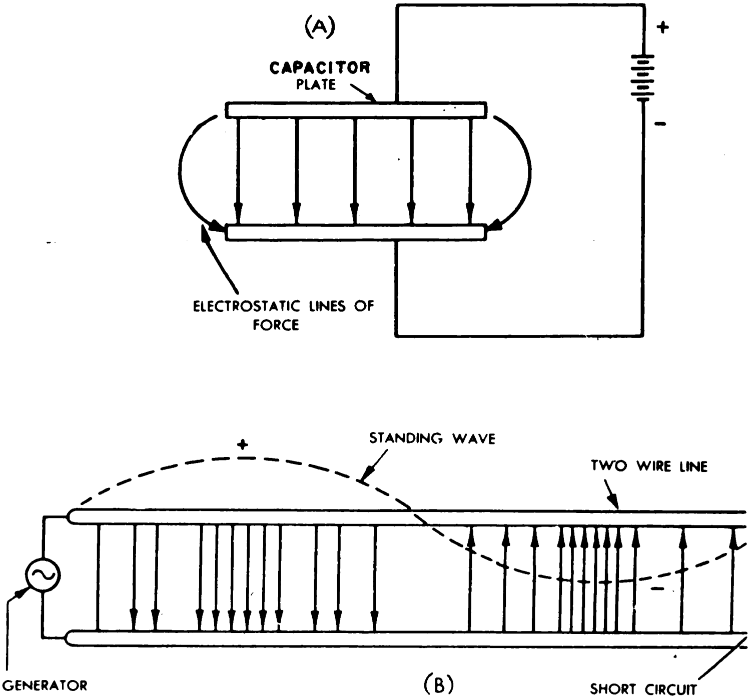


Figure 5–6.—Electric field between capacitor plates and a full-wave section of two-wire line.

arrows. A long arrow represents more stress than a short arrow. In representing electrostatic fields, the NUMBER of arrows indicates the strength of the field (fig. 5–6). In the case of the capacitor, note that the arrows are evenly spaced across the area between the two plates. As the voltage across the plates is the same at all points, the electrostatic lines between the plates are evenly distributed. This set of lines forms an electrostatic field. This field is usually called the ELECTRIC FIELD and is abbreviated the E-FIELD. The lines of stress are called the E-LINES.

Notice in (B) of figure 5–6 that the two-wire transmission line has an instantaneous standing wave of voltage applied to it. This line is equal to one wavelength. At the same time, part of it is positive, while another part is negative. The instantaneous electrostatic field (E-field) is the same at the negative and positive points, but the arrows representing each field point in opposite directions. The voltage along the line varies sinusoidally. Therefore, the density of the E-lines varies sinusoidally.

An easy way to show the development of the E-field in a waveguide is from a two-wire line which has quarter-wave insulators. Figure 5–7 shows the two-wire line previously discussed with several double quarter-wave insulators, or half-wave frames used as insulators.

The E-field on the main line is the same as that in the transmission line, as illustrated in figure 5–6. The half-wave frames located at points of high voltage (strong E-field) will have a strong E-field across them. The half-wave frames located at a voltage minimum point will have no E-field across them. Frame (A) of figure 5–7 is an example of an insulator which has a strong E-field across it. Each frame is shown separately below the main line for a clearer view. Frame (B) is at zero voltage point, so it will have no field on it. Frame (C) also has a strong field, but its polarity is reversed. Frame (D) has a weaker field on it due to its being at a lower voltage point on the main line. The picture shown is

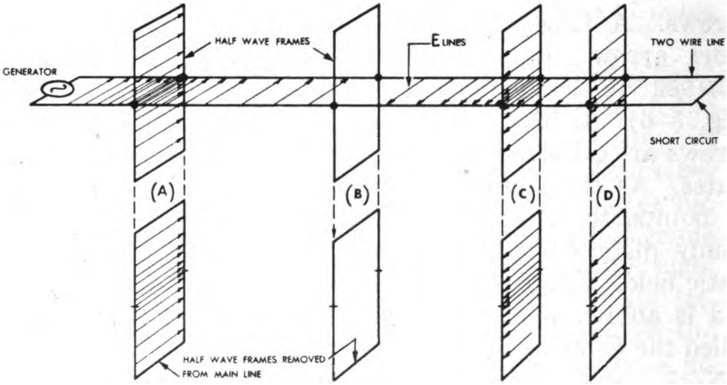


Figure 5–7.—Magnitude of fields on half-wave frames vary with strength of field on main line.

a buildup to the three dimensional aspect of the full E-field in a waveguide.

Figure 5–8 shows the E-field in an actual waveguide. This is the field which results when an infinite number of quarter-wave sections are connected to the line to form a rectangular box. The E-field is strong at one-quarter and three-quarter distances from the shorted end, but becomes weaker at the sine rate toward the upper and lower walls and toward the ends and center. Again the phenomenon of wavelength is present, as shown in (A). You should realize, of course, that this is an instantaneous picture taken at the time the standing wave of voltage is at its peak. At other times, the voltage and E-field vary from zero to the peak value, reversing direction every alternation of the applied voltage.

Certain boundary conditions must exist in order for propagation to occur in a waveguide. The principal one is that there must be no electric field tangent to the walls of the guide. This is satisfied by the E-field diminishing to zero at the top and bottom of the guide by natural RF line action; while at other places, the field is perpendicular to the walls.

THE MAGNETIC FIELD.—The second field which must always exist in a waveguide is the magnetic field. The

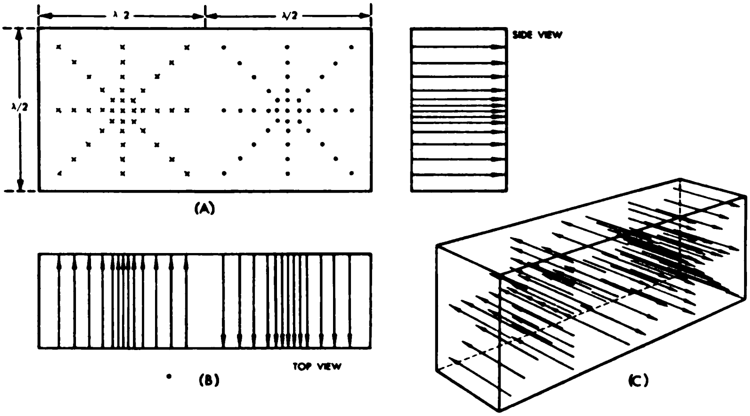
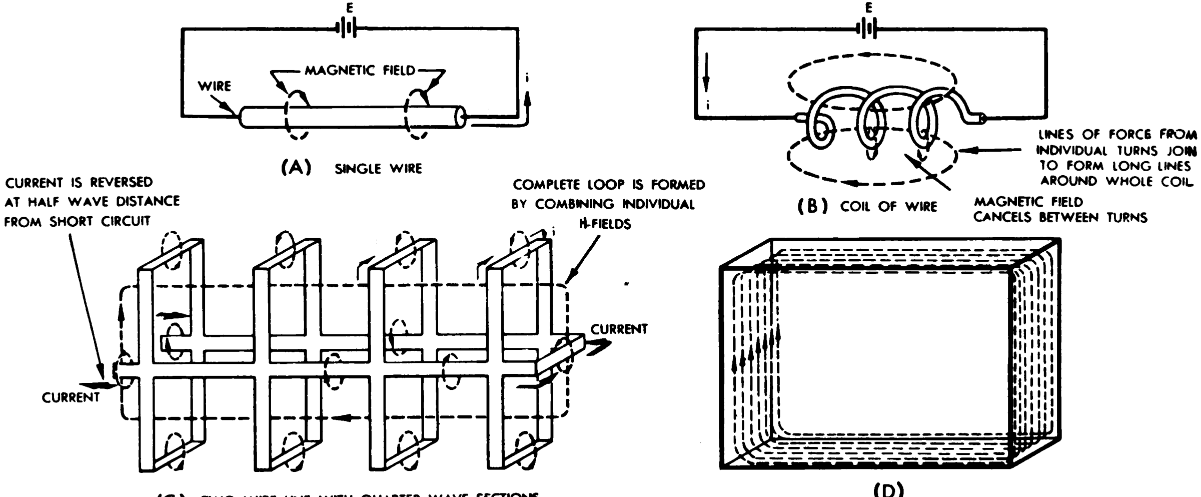


Figure 5–8.—E-field in an actual waveguide.

magnetic lines of force which make up the magnetic field are caused by the MOVEMENT of electrons in the conducting material. All the tiny magnetic forces exerted by the individual moving electrons add together and form a large force around the conductor. The presence of the force is shown by closed loops around the single wire in (A) of figure 5–9. The closed loops are formed of magnetic lines of force, or H-lines, which must be continuous, or closed, in order to exist. These lines are associated with a current; their number varies directly with the amount of current present; and the total number of lines considered collectively is called the H-FIELD. Each H-line has a definite direction, which can be determined by use of the left-hand rule; and in the diagrams which follow, the strength of the field is indicated by the number of H-lines drawn in a given area.

Although H-lines encircle a single straight wire, they behave differently when the wire is wound into a coil. (See fig. 5–9 (B).) In a coil the individual H-lines tend to form around each turn of wire, but in doing so take opposite directions between adjacent turns. This causes cancellation and results in zero field strength between the turns. But inside and outside the coil, the directions


E
MAGNETIC FIELD
WIRE
i
E
i
LINES OF FORCE FROM INDIVIDUAL TURNS JOIN TO FORM LONG LINES AROUND WHOLE COIL
MAGNETIC FIELD CANCELS BETWEEN TURNS
CURRENT IS REVERSED AT HALF WAVE DISTANCE FROM SHORT CIRCUIT
COMPLETE LOOP IS FORMED BY COMBINING INDIVIDUAL H-FIELDS
i
CURRENT
CURRENT

(A) SINGLE WIRE

(B) COIL OF WIRE

(C) TWO WIRE LINE WITH QUARTER WAVE SECTIONS

(D)

Figure 5–9.—Development of magnetic field or H-field in the waveguide.

are the same for each H-field. Therefore, the fields join and form a continuous H-line around the entire coil.

Similar action also takes place in a waveguide. In diagram (C) of figure 5–9, a two-wire line with quarter-wave sections is shown. Currents flow in the main line and in the quarter-wave sections. The current direction produces the individual H-lines around each conductor. When a large number of sections exist, the field cancels between the sections but their directions are the same inside and outside the waveguide. At half-wave intervals on the main line, current will flow in opposite directions. This produces H-line loops having opposite directions. In (C), current at the left end is opposite to the current on the right end. The individual loops on the main line are opposite in direction. All around the framework, they join in such a manner that the long loops shown in (D) are formed. Outside of the waveguide, the individual loops cannot join to form a continuous loop. Thus, there is no magnetic field outside of a waveguide.

Figure 5–10 shows a conventional presentation of the magnetic field in a waveguide three half wavelengths long. Note that the field is strongest at the edges of the waveguide. This is where the current is the highest. The current is lowest at the center of each set of loops because there the standing wave of current is zero at all times. The picture shown represents an instantaneous condition. During the peak of the other half cycle of a-c input, all field directions are reversed. An instantaneous picture of this condition would show the fields reversed at half-wave intervals, since the current in the long line is reversed over half-wave distances.

A second boundary condition necessary for electromagnetic fields to transfer power is satisfied by the configuration of the magnetic field. This condition requires that at the surface of the waveguide there be no perpendicular component of the magnetic field. Since all the H-lines are parallel to the surface, this condition is satisfied. Thus, energy can be transmitted by a waveguide

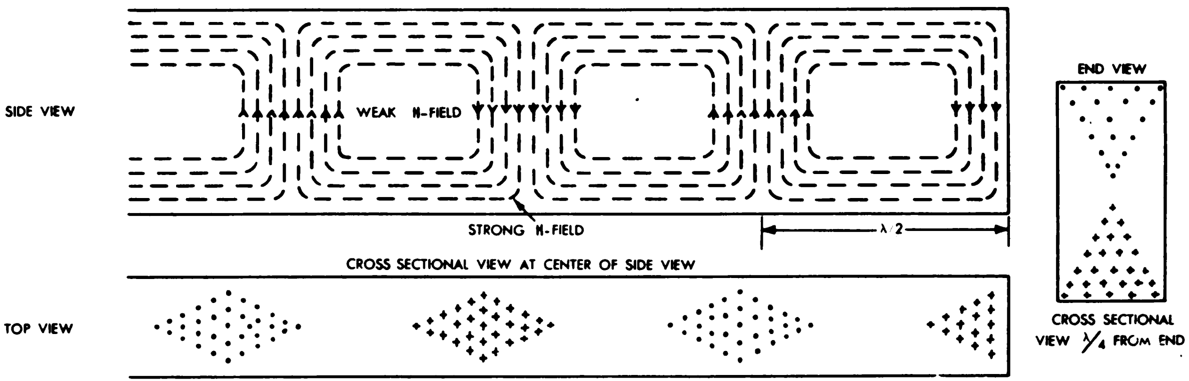

SIDE VIEW
WEAK H-FIELD
STRONG H-FIELD
λ/2
END VIEW
CROSS SECTIONAL VIEW AT CENTER OF SIDE VIEW
TOP VIEW
CROSS SECTIONAL VIEW λ/4 FROM END

Figure 5–10.—Magnetic field in waveguide three half wavelengths long.

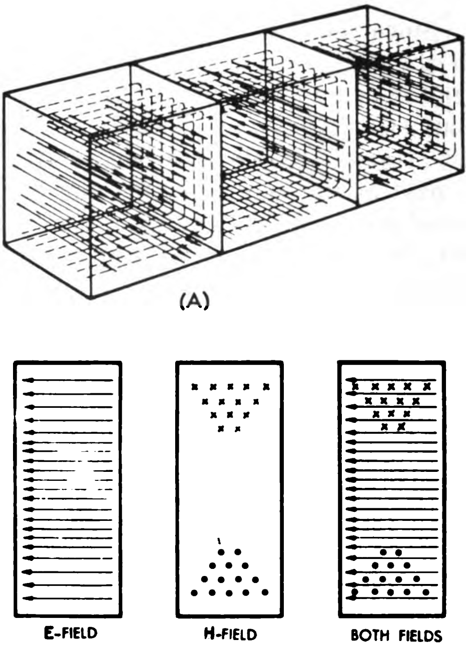
and will be carried in the electromagnetic fields within the guide. Since the electric field is zero wherever it is parallel to the metal surface of the guide, no part of this field can penetrate the metal. In the same way, the varying magnetic field is zero at the surface of the metal wherever it attempts to penetrate the walls of the guide, so that no part of the magnetic field can get outside the guide. Therefore, all of the energy in the field is wholly contained by the guide. Thus, measurements on the outside of the guide can detect neither oscillating electrons nor any electromagnetic fields caused by the electrical energy within the waveguide.

Electric and magnetic fields exist simultaneously in the waveguide. In fact, the H-field causes a current which in turn causes a voltage difference. This causes an E-field and it in turn causes a current which causes an H-field. This is a continuous action. One field is dependent on the other as energy is continually transferred from one field to the other.

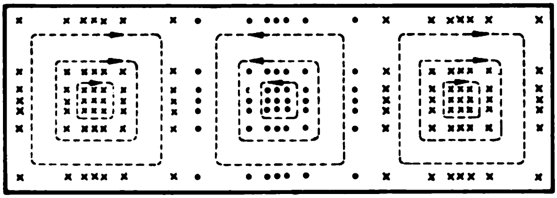
The conventional picture of both fields in the waveguide is shown in (A) of figure 5–11. Since this picture is rather complicated, the presence and direction of the field is usually indicated in simpler diagrams, such as (B), (C), and (D). In these diagrams the number of E-lines in a given area indicates the strength of the electrostatic field, while the number of H-lines in any given cross section indicates the strength of the magnetic field in that area.

The field configuration shown in this illustration represents only one of the many ways in which fields are able to exist in a waveguide. Such a field configuration is called a MODE of operation. In the case of the rectangular waveguide illustrated in figure 5–11 the configuration is known as the DOMINANT mode, since it is the easiest one to produce. Other higher modes—that is, different field configurations—may occur accidentally or may be caused deliberately in a waveguide.

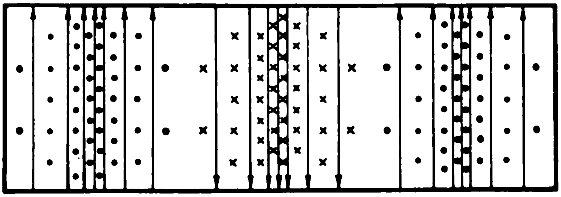
An example of another field configuration is developed in (A) of figure 5–12. If the size of this waveguide is



(B) END VIEWS



(C) SIDE VIEW



(D) TOP VIEW

Figure 5–11.—Conventional picture of both fields in a waveguide.

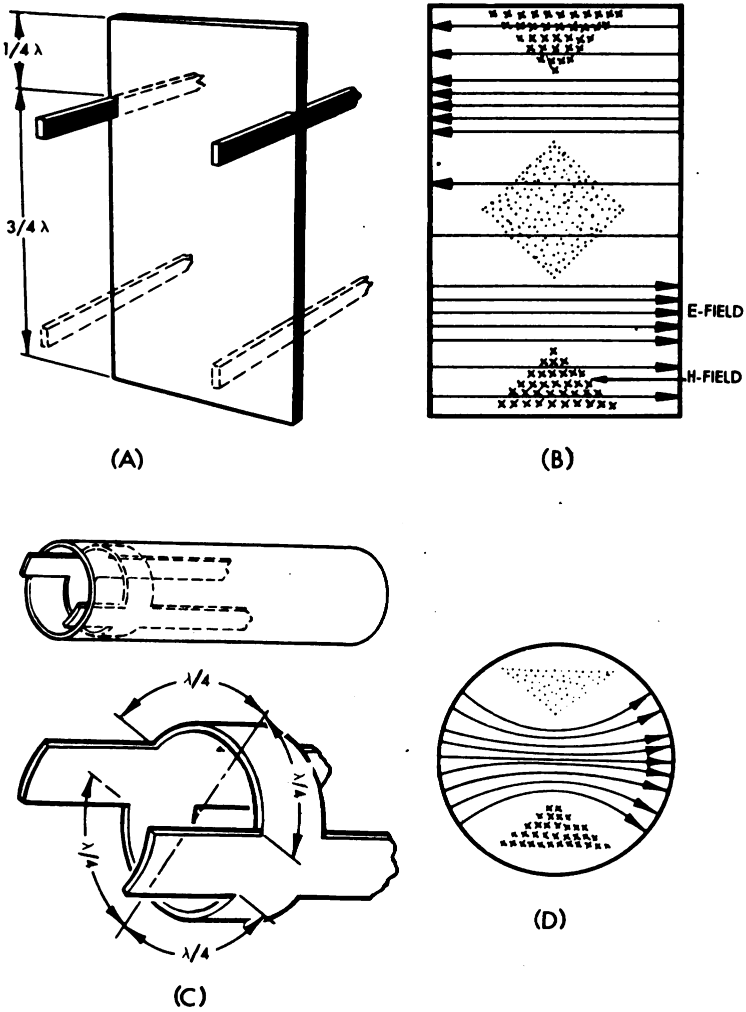


Figure 5–12.—Other field configurations in waveguides.

doubled over that of the waveguide shown in the previous illustration, the cross section will be a full wave rather than a half wave. The two-wire conductor can be assumed to be a quarter wave down from the top (or a quarter wave up from the bottom). The remaining distance to the bottom is three-quarters of a wave. A three-quarter wave section has the same high impedance input as the quarter-wave section. Thus the two-wire line is properly insulated and will transfer energy. The field configuration will show a full wave across the wide dimension, as you can see in (B).

This field configuration can be applied to a circular waveguide. The two conductors shown in (C) of figure 5–12 are assumed to be part of the waveguide wall. The remaining part of the wall forms the quarter-wave sections each of which acts as an insulator between the two conductors. This makes it possible to transfer energy with minimum losses. The resulting field configuration, shown at (D), is the dominant mode for a waveguide with a circular cross section.

Analogy of Waveguide Action by Electric Waves

A somewhat different analogy involving waveguide action deals with the field rather than with current and voltages. This analogy is somewhat more exact than the previous explanation which dealt with voltages and current in two-wire lines, for power flow in this case is assumed to be in the fields rather than in the conductors.

In a waveguide the fields are the same as those radiated into space by an antenna. Figure 5–13 shows a small portion of such a field which is radiated into space from an antenna. In it, the electrostatic lines of force, or E-lines, are parallel to the antenna, and the electromagnetic lines of force or H-lines are perpendicular to the antenna. They move away from the antenna at the speed of light. At each half cycle the polarity is reversed. Therefore, at half-wave intervals, the fields are in opposite direction (or polarity). Although only a small part of the total field is shown, actually the E-lines and H-lines form huge closed loops after they leave the antenna.

As previously mentioned, the energy which moves through a waveguide and the energy which is radiated by an antenna are both the same form of electromagnetic

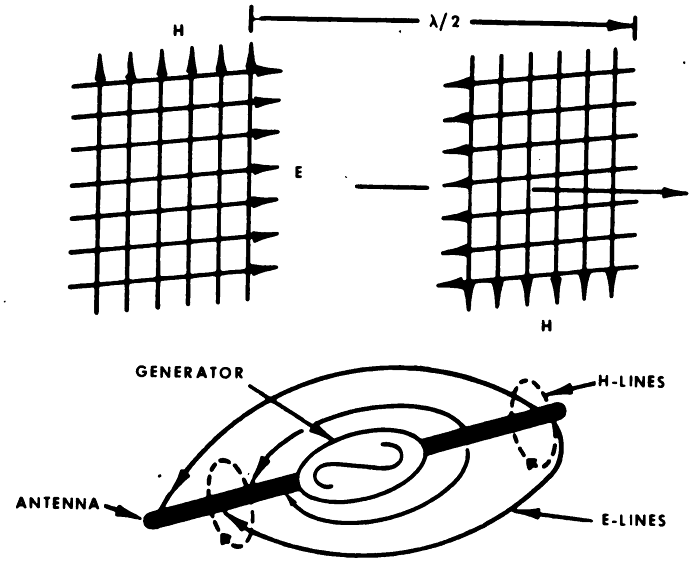


Figure 5–13.—Small portion of field radiated into space by an antenna.

radiation. Nevertheless, the field configuration shown in figure 5–14 cannot exist in a waveguide because it does not satisfy the required boundary conditions. First, there cannot be any E-lines tangent to the surface of the walls. Since the E-lines are evenly distributed across the area, some will be across the top and bottom walls. This causes the voltage to short out, and, in turn, the E-lines vanish. Other E-lines which are pushed up to the wall by the repulsion between E-lines likewise short out. This shorting out is cumulative and eventually removes the entire E-field.

The second boundary condition which must be satisfied is that there must be no component of the magnetic field perpendicular to the wall. Note again in the illustration that the H-lines are parallel to the side walls which is correct—but are perpendicular to the bottom, which cannot be; so H-lines of this type cannot exist in the waveguide either. Furthermore, an H-line cannot exist unless it is a closed loop.

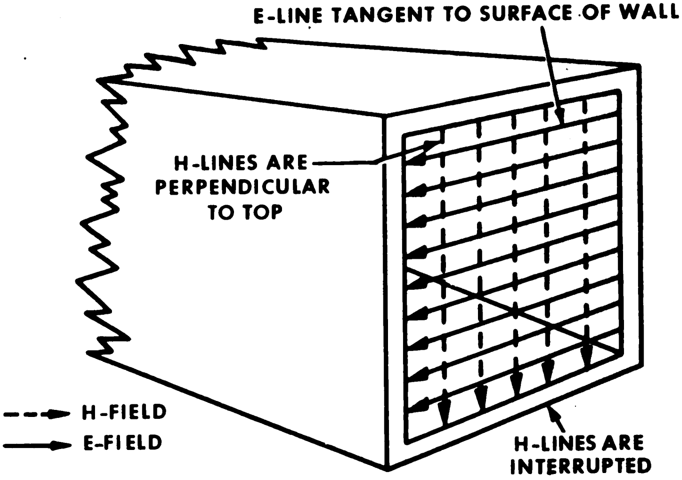

E-LINE TANGENT TO SURFACE OF WALL
H-LINES ARE PERPENDICULAR TO TOP
H-FIELD
E-FIELD
H-LINES ARE INTERRUPTED

Figure 5–14.—Fields in a waveguide must satisfy boundary conditions to be radiated.

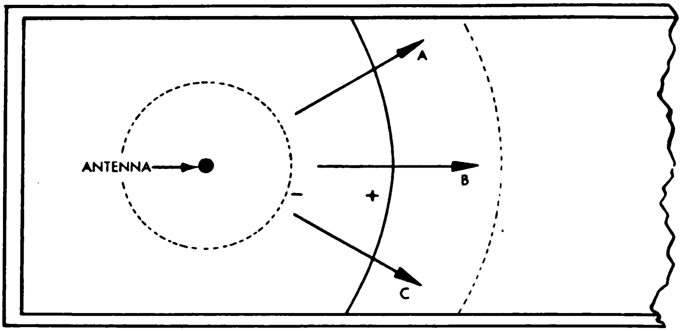


Figure 5–15.—How radiation fields are made to fit a hollow pipe.

When a small antenna is placed in the waveguide and excited at an RF frequency, both positive and negative half cycles are radiated. (See fig. 5–15.) The wavefront produced is like an expanding circle. The part which travels in the direction of arrow *B* goes straight down the waveguide and is quickly attenuated, as previously described. However, the part of the wavefront which travels in the direction of arrow *A* is reflected from the wall. The wall is a short circuit and causes the wavefront to be reflected in reverse phase. Meanwhile, the wavefront which travels in a direction *C* is reflected from the other wall and proceeds in opposite phase. Thus, the radiation fields are contained in the waveguide.

PATH OF WAVEFRONTS IN A WAVEGUIDE.—In the side view of the waveguide, figure 5–16 (A), the light solid and light broken lines represent the wavefront going in direction *A*. The heavy solid and heavy broken lines represent the wavefront going in direction *B*. Note that all parts of the wavefront *A* are traveling upward at an angle across the guide. Wavefront *B* is traveling at the same angle but downward.

When the wave travels in this fashion in a waveguide, propagation is possible. In understanding what happens, note that the positive wavefront (represented by both light and heavy SOLID lines) occurs simultaneously throughout the center of the guide. These fronts add and cause a maximum voltage to occur at the center. The E-field is shown maximum at the center in diagram (C) of figure 5–16. The negative wavefront (represented by both light and heavy BROKEN lines) adds in the same manner as the positive wavefront. When the negative wavefront meets the positive wavefront at the walls, the two wavefronts cancel each other, making the total voltage equal to zero. This verifies the E-field condition shown in (C). With the E-field zero at the edges, it is possible for the E-field to exist in the waveguide.

CROSSING ANGLE.—The angle at which a wavefront crosses a waveguide is a function of the wavelength and the cross-sectional dimension of the waveguide. At some intermediate frequency the reflection is as shown in (B) of figure 5–17. But as the frequency increases, the angle of incidence becomes less and the signal travels farther before it reaches the other side as shown in (A). At lower frequencies, the wavefront crosses the guide at nearly right angles to the walls. At some frequency, the angle will be 90 degrees. At this point, the wave travels back and forth across the guide until the energy is dissipated by the resistance of the walls of the guide. At

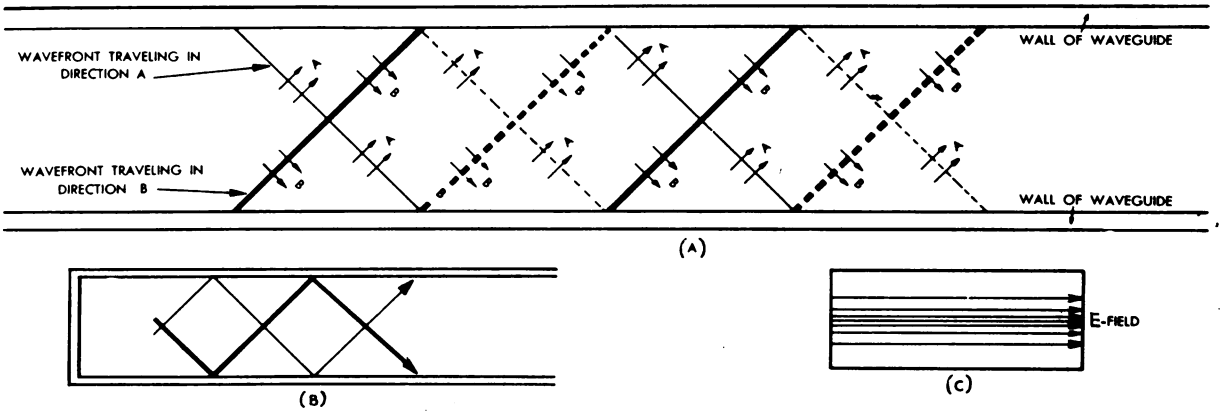

WAVEFRONT TRAVELING IN DIRECTION A
WAVEFRONT TRAVELING IN DIRECTION B
WALL OF WAVEGUIDE
WALL OF WAVEGUIDE
A
B
(A)
(B)
E-FIELD
(C)

Figure 5–16.—Paths of wavefronts in waveguide.

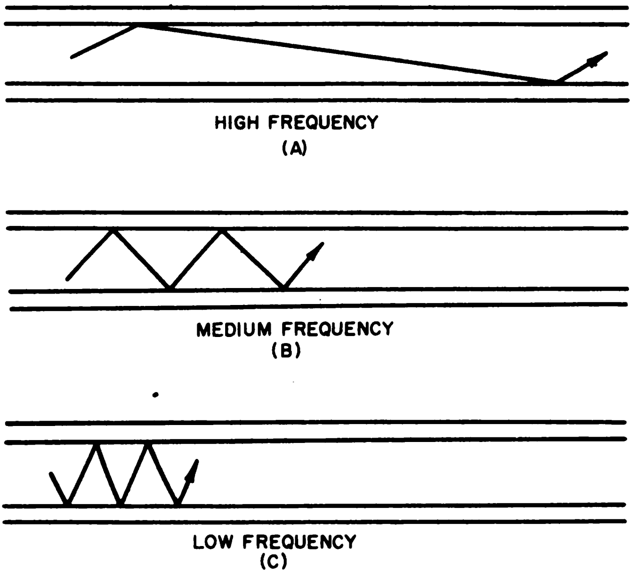


Figure 5–17.—Angle at which fields cross waveguide varies with frequency.

this frequency, the distance from side to side is one-half wavelength for the waveguide. At the cutoff frequency, the attenuation is a linear function of length and is very high.

The velocity of propagation of a wave along a two-wire line is less than its velocity in air. The same is true in a waveguide. Movement of a wave along a two-wire line is slower than its movement in air because of the retarding effect of the d-c resistance, the conductors, and conductance of the insulation. In the waveguide, the lower velocity is due to the way the field travels. As shown in (C) of figure 5–17, the path of a wavefront at a relatively low frequency is along the zigzag arrow at the velocity of light. But due to the long path, the wavefront actually travels very slowly along the waveguide. In figure 5–17 (A), the frequency is higher, and the wavefront or the group of waves actually travel a given distance in less time than those in (C).

The axial velocity of a wavefront or a group of waves

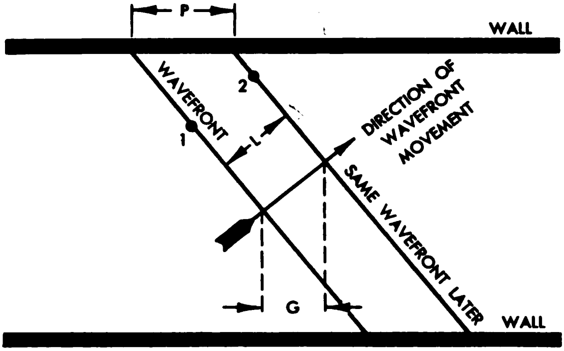


Figure 5–18.—Relation of phase, group, and wavefront velocity.

is called the GROUP VELOCITY. The relationship of the group velocity to diagonal velocity causes an unusual phenomenon. The velocity of propagation appears to be greater than the speed of light. As you can see in figure 5–18, during a given time a wavefront will move from point 1 to point 2, or a distance *L* at the velocity of light (V_L). Due to this diagonal movement (direction of the arrow), the wavefront during this time has actually moved down the guide only the distance *G*, which is necessarily a lower velocity. This is called group velocity (V_g). But if an instrument were used to detect the two positions at the wall, they would be the distance *P* apart. This is greater than the distance *L* or *G*. The movement of the contact point between the wave and the wall is at a greater velocity. Since the phase of the RF has changed over the distance *P*, this velocity is called the phase velocity (V_p). The mathematical relationship between the three velocities is stated by the equation

$$V_L = \sqrt{V_p V_g}$$

where

V_L = velocity of light = 3×10^8 meters/second

V_p = phase velocity

V_g = group velocity

This equation indicates that it is possible for the phase velocity to be greater than the velocity of light. As the frequency decreases, the angle of crossing is nearer a right angle. In this condition the phase velocity increases. For measuring standing waves in a waveguide, it is the phase velocity which determines the distance between voltage maximum and minimum. For this reason, the wavelength measured in the guide will actually be greater than the wavelength in free space. From a practical standpoint, the different velocities are related in the following manner: if the RF frequency being propogated is sine-wave modulated, the modulation envelope will move forward through the waveguide at the group velocity, while the individual cycles of RF energy will move forward through the modulation envelope at the phase velocity. If the modulation is a square wave, as in radar transmissions, again the square wave will travel at group velocity, while the RF waveshape will move forward within the envelope. Since the standing wave measuring equipment is affected by each RF cycle, the wavelength will be governed by the rapid movement of the changes in RF voltage. Since intelligence is conveyed by the modulation, the transfer of intelligence through the waveguide will be slower than the speed of light, as is the case in other types of RF lines.

Because of the way the fields are assumed to move across the waveguide, it is possible to establish a number of trigonometric relationships between certain factors. As shown in figure 5–19, the angle that the wavefront makes with the wall (angle Θ) is related to the wavelength and dimension of the guide and is equal to

$$\text{Cos } \theta = \frac{\lambda}{2B}$$

where λ is the wavelength in free space of the signal in the guide, and B is the inside wide dimension of the guide. The group velocity (V_g) is related to the velocity of light (V_L) as follows:

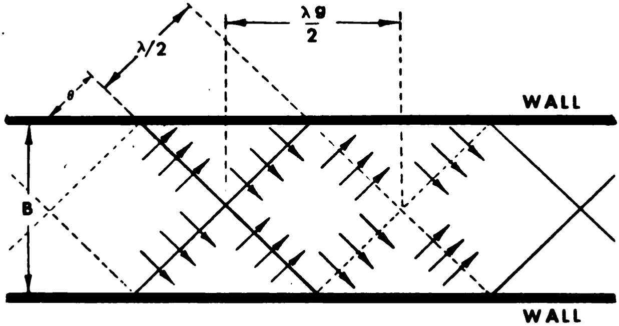


Figure 5–19.—Trigonometric relations exist between factors indicated.

$$\frac{V_g}{V_L} = \sin\theta = \sqrt{1 - \frac{(\lambda)^2}{(2B)^2}}$$

Further, since it is possible to measure the wavelength in the guide (λ_g), the wavelength in space is equal to

$$\frac{\lambda_g}{\lambda} = \frac{1}{\sin\ \theta} = \frac{1}{\sqrt{\frac{1-(\lambda)^2}{(2B)^2}}}$$

Solving for λ, the equation becomes equal to

$$\lambda = \frac{2B\lambda_g}{\sqrt{\lambda_g{}^2 + 4B^2}}$$

After measuring the wavelength and the inside dimension of the waveguide, it is possible to calculate most other quantities associated with the waveguide.

NUMBERING SYSTEM OF THE MODES.—The normal configuration of the electromagnetic field within a waveguide is called the DOMINANT mode of operation. The mode developed for the rectangular waveguide, as was explained before and illustrated in figure 5–10, is the dominant mode of operation. The dominant mode for the circular waveguide was also shown in figure 5–12 (D). A wide variety of higher modes are possible in either type of waveguide. The higher modes in the rectangular waveguides are seldom used in radar, but some of the higher modes in the circular waveguide are useful.

For ease in identifying modes, any field configuration can be classified as either a TRANSVERSE ELECTRIC MODE or a TRANSVERSE MAGNETIC MODE. These modes are abbreviated *TE* or *TM* respectively. In a transverse electric mode, all parts of the electric field are perpendicular to the length of the guide and no E-line is parallel to the direction of propagation. The *TE* mode is sometimes called the H-mode. In a transverse magnetic mode, the plane of the H-field is perpendicular to the length of the waveguide. No H-line is parallel to the direction of propagation. This mode is sometimes called an E-mode.

It is interesting to note from these definitions that the wavefront in free space or in a coaxial line is a *TEM* mode, since both fields are perpendicular to the direction of propagation. This mode cannot exist in a waveguide.

In addition to the letters *TE* or *TM*, subscript numbers are used to complete the description of the field pattern. In describing field configurations in rectangular guides, the first small number indicates the number of half-wave patterns of the transverse lines which exist along the short dimension of the guide through the center of the cross section. The second small number indicates the number of transverse half-wave patterns that exist along the long dimension of the guide through the center of the cross section. For circular waveguides the first number indicates the number of full waves of the transverse field encountered around the circumference of the guide. The second number indicates the number of half-wave patterns that exist across the diameter.

COUNTING WAVELENGTHS FOR MEASURING MODES.—In the rectangular mode shown in figure 5–20 (A), note that all the electric lines are perpendicular to the direction of movement. This makes it a *TE* mode. In the direction across the narrow dimension of the guide parallel to the E-line, the intensity change is zero. Across the guide along the wide dimension, the E-field varies from zero at the top, through maximum at the center, to zero on the bottom. Since this is one-half wave, the

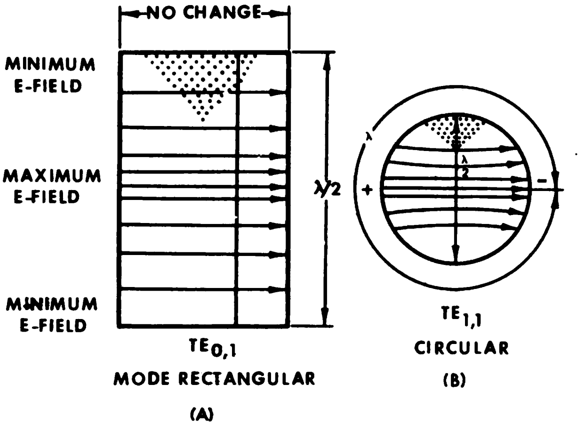


Figure 5–20.—How to count wavelengths for numbering modes.

second subscript is one. Thus, the complete description of this mode is $TE_{0,1}$.

In the circular waveguide in (B), the E-field is transverse and the letters which describe it are *TE*. Moving around the circumference starting at the top, the field goes from zero, through maximum positive (tails of arrows), through zero, through maximum negative (head of arrows), to zero. This is one full wave, so the number is one. Going through the diameter, the start is from zero at the top wall, through maximum in the center to zero at the bottom, one-half wave. The second subscript is one. The complete designation for the circular mode becomes $TE_{1,1}$.

Several circular and rectangular modes are possible. On each diagram of figure 5–21 you can verify the numbering system.

The side view of the waveguides in figure 5–22 show that the magnetic and electric fields are maximum in intensity in the same area. This indicates that the current and voltage are in phase. This is the condition which exists when there are no reflections to cause standing waves. In previous examples in which fields were developed, the fields were out of phase because of a short circuit at the end of the two-wire line.

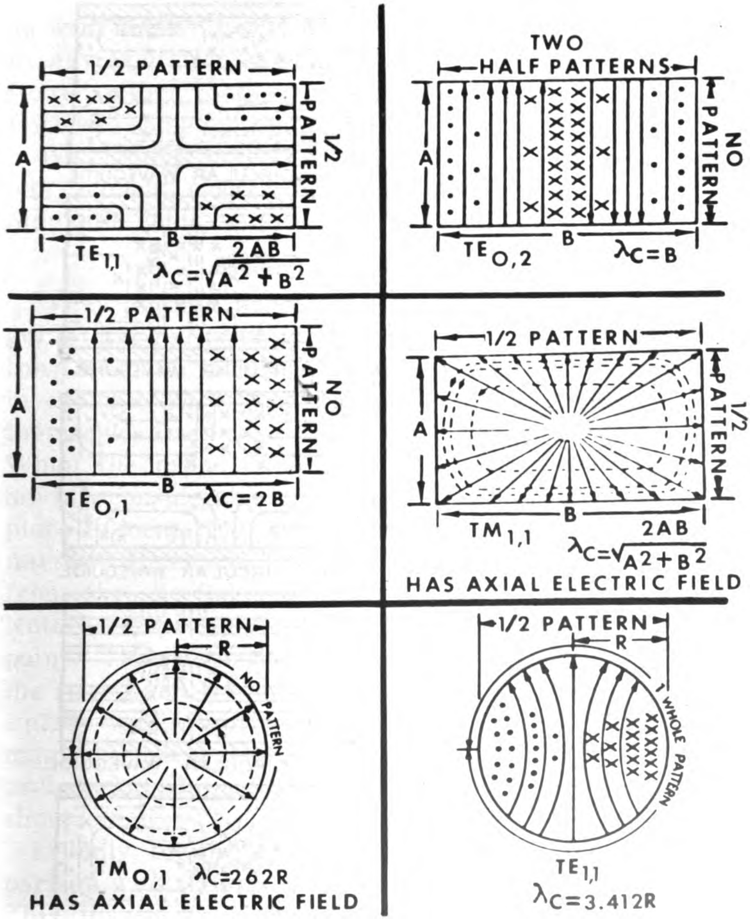


Figure 5–21.—Various modes in waveguides.

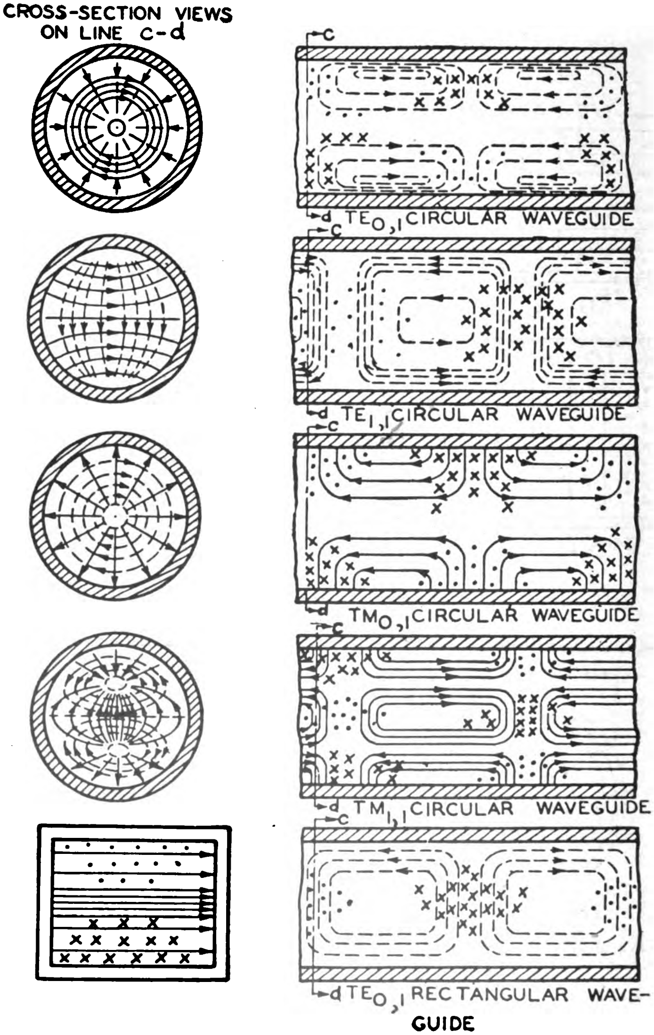

CROSS-SECTION VIEWS ON LINE c-d
c
d
$TE_{0,1}$ CIRCULAR WAVEGUIDE
c
d
$TE_{1,1}$ CIRCULAR WAVEGUIDE
c
d
$TM_{0,1}$ CIRCULAR WAVEGUIDE
c
d
$TM_{1,1}$ CIRCULAR WAVEGUIDE
c
d
$TE_{0,1}$ RECTANGULAR WAVE-GUIDE

Figure 5–22.—Various modes in waveguides.

INTRODUCING FIELDS INTO A WAVEGUIDE

A waveguide, as was explained before, is a single conductor. Therefore, it does not have the two connections which ordinary RF lines have, and it is necessary to use special devices to put energy into a waveguide at one end and to remove it from the other. In a waveguide, as with many other electrical networks, reciprocity exists in any excitation system—that is, energy may be transferred either to the waveguide or from the waveguide with the same efficiency.

Waveguides may be excited by three principal methods; namely, electric fields, magnetic fields, and electromagnetic fields.

Exciting With Electric Fields

When a small probe or antenna is placed in a waveguide and fed with an RF signal, current will flow in the probe and set up an electrostatic field such as shown in figure 5–23 (A). This causes the E-lines to detach themselves from the probe and to form in the waveguide. When the probe is located in the right place, a field having considerable intensity will be set up. The best place to locate the probe is in the center, parallel to the narrow dimension and one-quarter wavelength away from the shorted end of the guide, as shown in (C). Note here that the field is strongest at the quarter-wave point. This is the point of maximum coupling between the probe and the field. Of course, the probe will work equally well at any point where E-lines exist with maximum intensity. For example, a good spot to place the probe would be a three-quarter wave distance from the shorted end.

Usually, the probe is fed with a coaxial cable. In comparison with the waveguide, this cable is extremely short. This insures that the greatest benefit will be derived from the waveguide. Impedance matching between the coaxial cable and the waveguide is accomplished by varying the distance from the probe to the end of the waveguide (by moving the shorted end), and by varying the

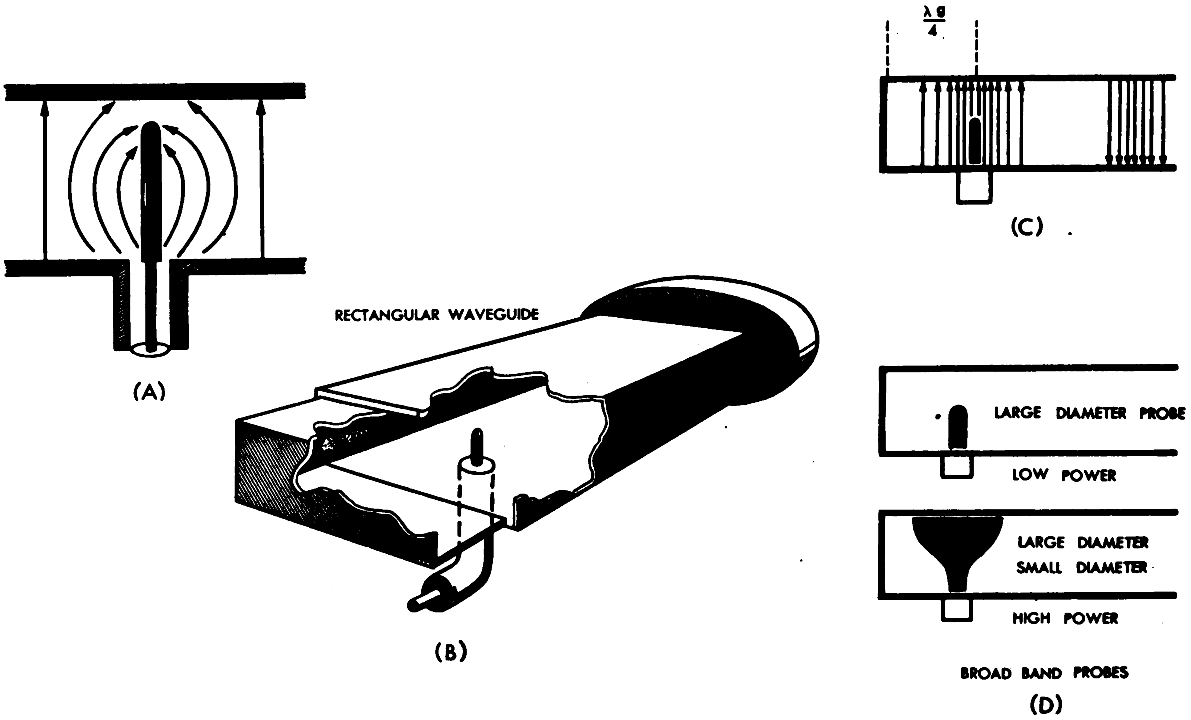


Figure 5–23.—Exciting the waveguide with electric field.

length of the probe (fig. 5–23 (B)). A mismatch will cause unwanted reflections in the waveguide.

The degree of excitation can be reduced by reducing the length of the probe, moving it out of the center of the E-field, or shielding it. Where it is necessary to vary the degree of excitation frequently, the probe is made retractable and the end of the waveguide fitted with a movable plunger. In airborne radar systems, the position of the probe and the end piece is often predetermined at the factory and fixed permanently.

In pulse-modulated radar systems there are wide sidebands on each side of the carrier. In order that a probe feeding system does not discriminate too sharply against frequencies which differ from the carrier frequencies, WIDE-BAND probes are often used. This probe is large in diameter and is conical or doorknob in shape. A conical probe is capable of handling high powered signals ((D) in fig. 5–23). The same kind of probe is used to take energy out of the guide and deliver it to the coaxial cable.

Exciting With a Magnetic Field

Another way of exciting a waveguide is by setting up a magnetic H-field in the waveguide. This can be accomplished by placing a small loop, which carries a high current, in the waveguide. This is what occurs: A magnetic field builds up and expands until it fills the space within the waveguide. If the frequency of the current is correct, energy will be transferred from the loop to the waveguide. A loop for transferring energy into a guide is shown in figure 5–24 (A) and (B). Notice that the loop is fed by a coaxial cable. The location of the loop for optimum coupling to the guide is at the place where the magnetic field, which is to be set up, is of greatest strength. There are a series of places where this is true. Several are shown in (C).

When less coupling is desired, you can rotate or move the loop until it encircles a smaller number of lines of force.

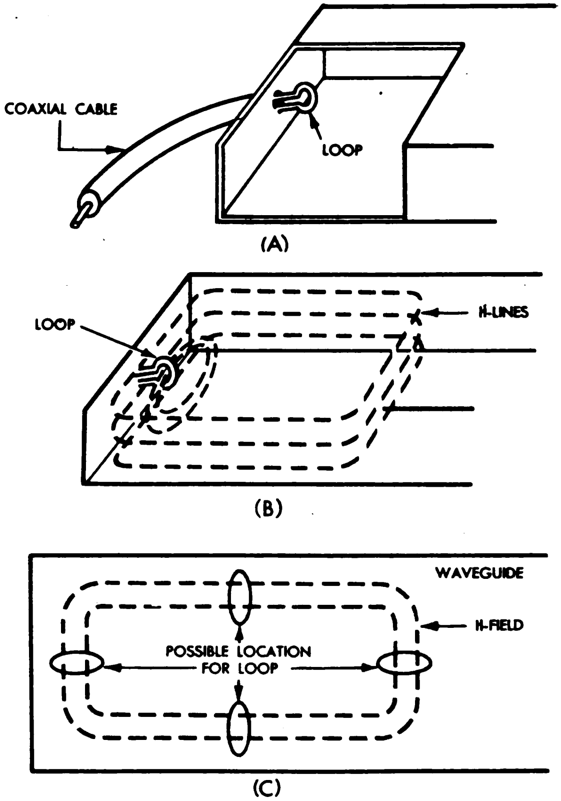


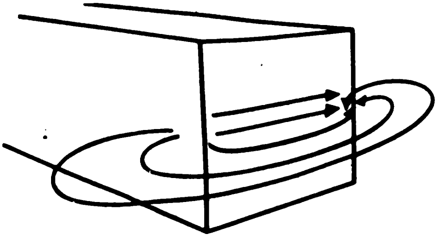
Figure 5–24.—Excitation with magnetic field.

When an excitation loop is used in radar equipment, its proper location is often predetermined and fixed either during construction or final tuning at the factory. In test or laboratory equipment, the loop is often made adjustable.

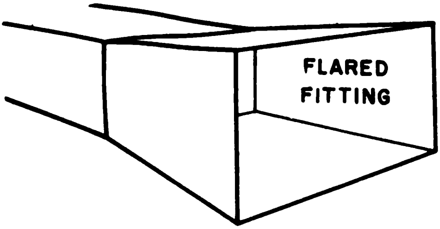
When a loop is introduced in a guide in which an H-field is present, a current will be induced in the loop itself. Thus, the loop may take energy out of the waveguide as well as put energy into it.

Exciting With Electromagnetic Fields

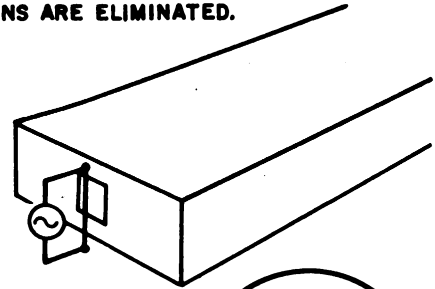
After learning what fields are like, you might think that a good way either to excite the waveguide or to couple energy out of it is simply to leave the end open. However, this is not the case, for when energy leaves a guide, fields exist around the end which result in mismatch, as shown in figure 5–25 (A). In other words,



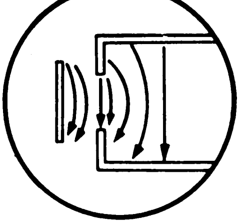
(A) REFLECTIONS OCCUR FROM AN ORDINARY OPEN END DUE TO THE WAY FIELDS EXPAND AROUND OPENING.



(B) BY FLARING OPEN END WITH OPTIMUM PROPORTIONS, REFLECTIONS ARE ELIMINATED.



(C) EXCITATION THROUGH APERATURE.



(D) FIELDS LEAK THROUGH APERATURE.

Figure 5–25.—Excitation with electromagnetic fields.

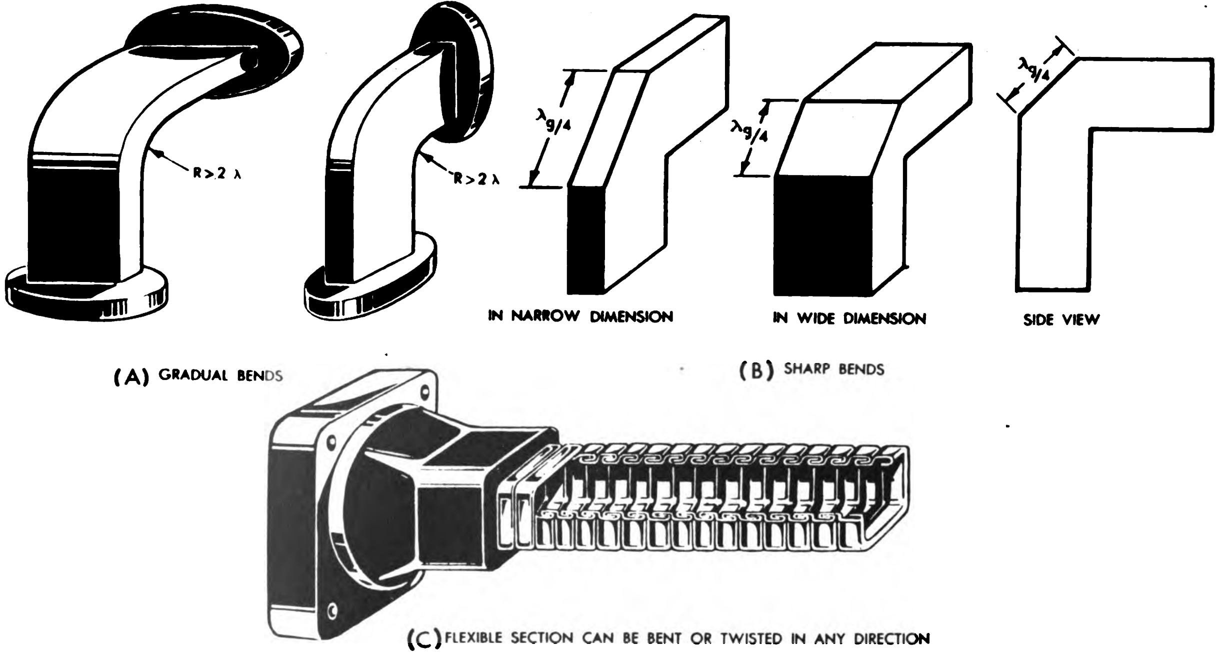
reflections and standing waves would result if the end were left open. Thus, simply leaving the end open is not an efficient way of letting energy out of the waveguide.

In order for energy to move smoothly in or out of a guide, the opening of the guide may be flared like a funnel, as shown in (B) of figure 5–25. This makes the guide similar to a V-type antenna. The funnel in effect eliminates reflection by matching the impedance of free space to the impedance of the waveguide. When the mouth of the funnel is exposed to electromagnetic fields, they enter and are gradually shaped to fit the waveguide. The funnel is directional in characteristic. It sends or receives the greatest amount of energy in front of the opening.

Another method for either putting energy into or removing it from waveguides is through slots or openings. This method is sometimes used when very loose coupling is desired. In this method energy enters the guide through a small aperture, as you can see in (C). Any device which will generate an E-field may be placed near the aperture and the E-field will expand into the waveguide. A single wire is shown in (D). On it E-lines are set up parallel to the wire due to the voltage difference between parts of the wire. The E-lines, in expanding, will exist first across the aperture, then across the interior of the waveguide. If the frequency is correct and the size of the aperture properly proportioned, energy will be transferred to the waveguide with a minimum of reflections.

BENDS, TWISTS, JOINTS, AND TERMINATIONS

In order for energy to move from one end of a waveguide to the other without reflections, the size, shape, and dielectric material of the waveguide must be constant throughout its entire length. Any abrupt change in its size or shape results in reflections. Therefore, if no reflections are desired, any change in the direction or the size of the waveguide must be gradual. When it is necessary that the change in direction or size be abrupt,



(A) GRADUAL BENDS

(B) SHARP BENDS

(C) FLEXIBLE SECTION CAN BE BENT OR TWISTED IN ANY DIRECTION

Figure 5–26.—Types of bends.

then special devices, such as bends, twists, joints, or terminations, must be used.

Bends

Waveguides may be bent in several ways to avoid reflections (fig. 5–26). One is to make the bend gradual. It must have a radius of bend greater than two wavelengths in order to minimize any reflection. Some bends may be 90-degree bends; others may be greater or less than 90 degrees, depending upon the requirements of the system. Still another type of bend is the sharp bend.

A bend can be made in either the narrow or wide dimension of a guide without changing the mode of operation. Normally, in a sharp 90-degree bend, reflections will occur. To avoid this, the guide is bent twice at 45 degrees, one quarter wave apart. The combination of the direct reflection at one bend and the inverted reflection from the other bend will cancel and leave the fields as though no reflection had occurred.

To permit using any special bend which an installation might require, sections of a waveguide are often made flexible. These sections can be bent or twisted in any desired direction. They usually consist of a spiral wound ribbon of brass. In cross section the winding is exactly the same size as the waveguide. The entire assembly is like a spiral spring in that it can be bent or twisted into any desired shape. As skin effect keeps the current at the inner surface of the waveguide, the inside surfaces of the flexible section are chromium plated. This provides for maximum current conductivity. The outside of the section is covered with rubber. This gives the section flexibility and at the same time makes it both airtight and watertight.

Rotating the Field

Sometimes it is desired to rotate the electromagnetic fields so that they are in the proper direction for matching. This may be accomplished by twisting the waveguide (fig. 5–27). The twist should be gradual and

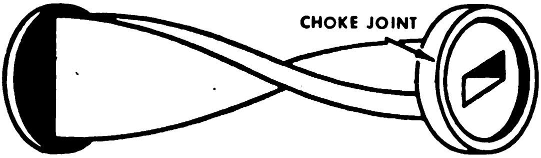


Figure 5–27.—Twisted section of waveguide rotates the field with minimum reflections.

extend over two wavelengths or more to prevent excessive reflections. Flexible sections also are used to rotate fields.

Joints

Since it is impossible to mold an entire waveguide system in a radar set into one piece, it is necessary to construct it in sections, and then to connect the sections together by joints. There are three main types of joints. These are the permanent, the semipermanent, and the rotating joints.

On the surface it would appear that joining two waveguide sections together would only require that the sections be the same size and fit tightly at the joint. However, irregularities at the joints set up standing waves and allow energy to escape. One kind of joint which affords a good connection between the parts of a waveguide and which has very little effect on the fields is the PERMANENT TYPE. This joint is made at the factory. When it is used, the waveguide sections are machined within a few thousandths of an inch and then welded together. The result is a hermetically-sealed and mirror-smooth joint.

Where it is necessary that sections be taken apart for normal maintenance and repair, it is impractical to use a permanent joint. To permit portions of the waveguide to be taken apart, they are commonly connected together with semipermanent joints. The most common type of semipermanent joint is the choke joint. A cross-sectional view of a choke joint is shown in figure 5–28 (A). It consists of two flanges which are connected to the wave-

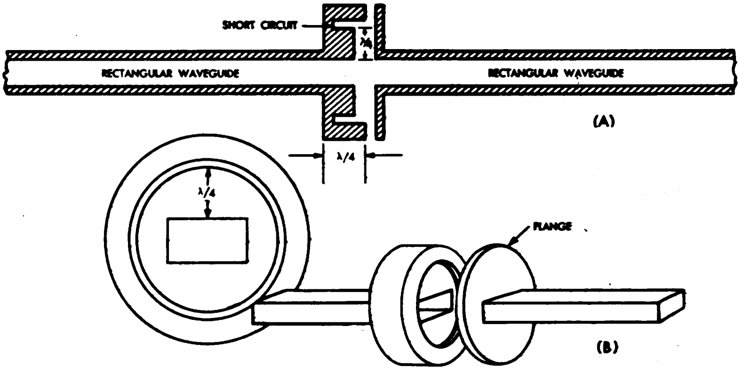


Figure 5–28.—Choke joints keep RF fields inside waveguide.

guide at the center. The right-hand flange is flat, and the one at the left is slotted a quarter wave deep at a distance a quarter wave from the point where the walls of the guide are joined. The quarter-wave slot is shorted at the end. The two quarter waves together become a half wave and reflect a short circuit at the place where the walls are joined together. Electrically, this creates a short circuit at the junction of the two waveguides. The two sections actually can be separated as much as a tenth of a wavelength without excessive loss of energy at this joint. This separation allows room to seal the interior of the waveguide with a rubber gasket for pressurization. The quarter-wave distance from the walls to the slot is modified slightly to compensate for the slight reactance introduced by the short space and the open circuit from the slot to the surface of the flange.

The name CHOKE JOINT is said to come from the similarity between the action of this joint on RF fields and the action of an RF choke in a power supply lead. An RF choke keeps RF in the circuit where it belongs. Similarly, the choke joint keeps the electromagnetic fields in the waveguide where they belong. The loss introduced by the well-designed choke is less than 0.03 db, while a well-machined, unsoldered permanent joint has a loss of 0.05 db or more.

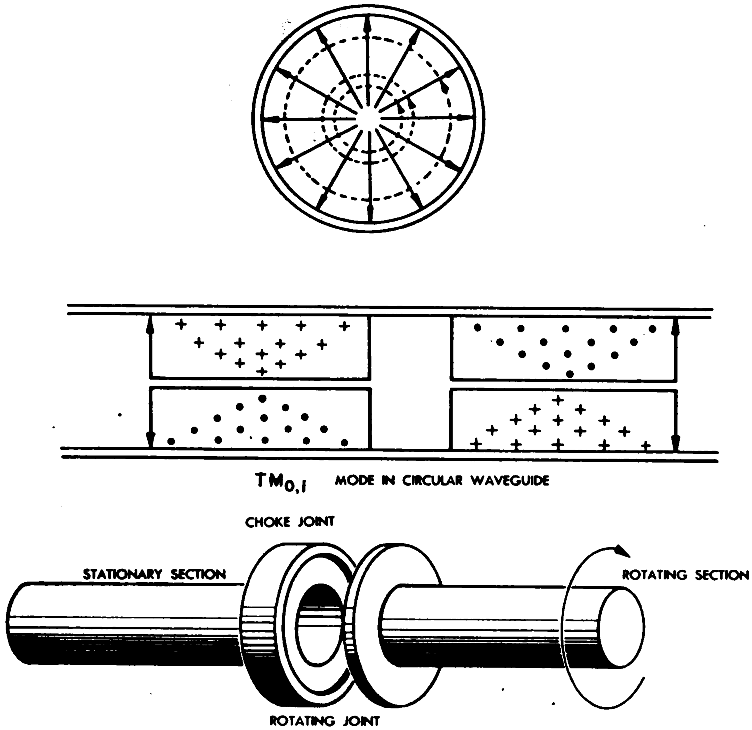


Figure 5–29.—Rotating joint and $TM_{0,1}$ mode in circular waveguide.

Rotating joints are usually required in an airborne radar system where the transmitter is stationary and the antenna is rotatable. A simple method for rotating part of a waveguide system is by using a mode of operation that is symmetrical about the axis (fig. 5–29). This requirement is met by using a circular waveguide and a mode such as $TM_{0,1}$. In this method a choke joint may be used to separate the sections mechanically and to join them electrically. With this method the waveguide rotates, but not the field. The continuous field produced by the fixed field minimizes reflections. Although this method is satisfactory, current airborne equipment uses rectangular waveguides. This requires a different system for connecting the sections of the waveguide. For me-

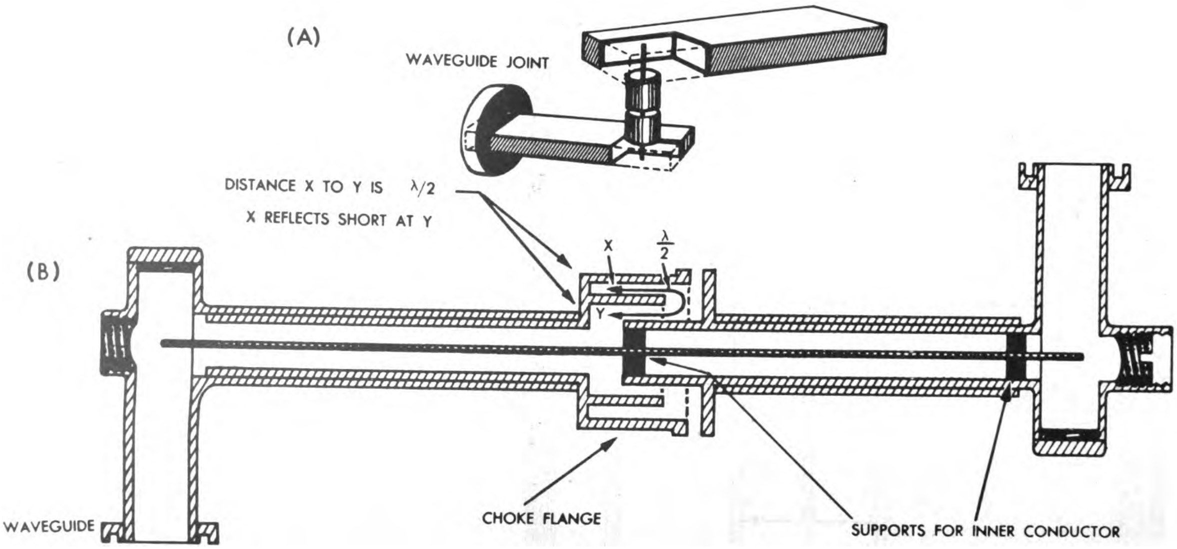


Figure 5–30.—Rotating joint with rectangular waveguide.

chanical reasons the rotating joint must be circular and involves the use of a coaxial cable, as shown in figure 5–30. This cable provides axial symmetry of the fields and the circular cross section for rotation. In the rotating joint, a probe forms the end of the center conductor of the coaxial cable. The probe takes energy from one waveguide. It is then conducted through the coaxial cable and delivered through another probe to the other waveguide.

The probe in the other waveguide is one end of the center conductor. The center conductor remains stationary with respect to one waveguide and rotates with respect to the other. To make the rotating electrical connection, the outer conductor can either be fitted with sliding contacts or with the tubing by a half-wave slot. The slot which is shorted at the end reflects a short at the junction of the two outer conductors. In this method, no mechanical contact is required between the two sections of the outer conductor. The inner conductor of the coaxial cable may be supported by insulating washers, if such are needed.

T-Junctions

Sometimes it is desirable to connect a section of waveguide into the side of another waveguide. This type of connection forms a T-junction. It may be connected either in the narrow side, as shown in figure 5–31 (A), or in the wide side of the waveguide, as shown in (D). When the T-junction is in the plane of an H-field of a $TE_{0,1}$ mode, it is called an H-TYPE junction, and when the junction is in the plane of the E-lines, it is called an E-TYPE junction.

The H-type junction is effectively a parallel connection with the main line. For example, when the end of the T-joint, (B) of figure 5–31, is short circuited at a distance of a half wave from the center of the waveguide, the result is the equivalent parallel circuit shown in (C). Note that (C) shows a half-wave section which is connected to a wire line. This section will reflect a short circuit at the line and will not allow any energy to pass. Similarly,

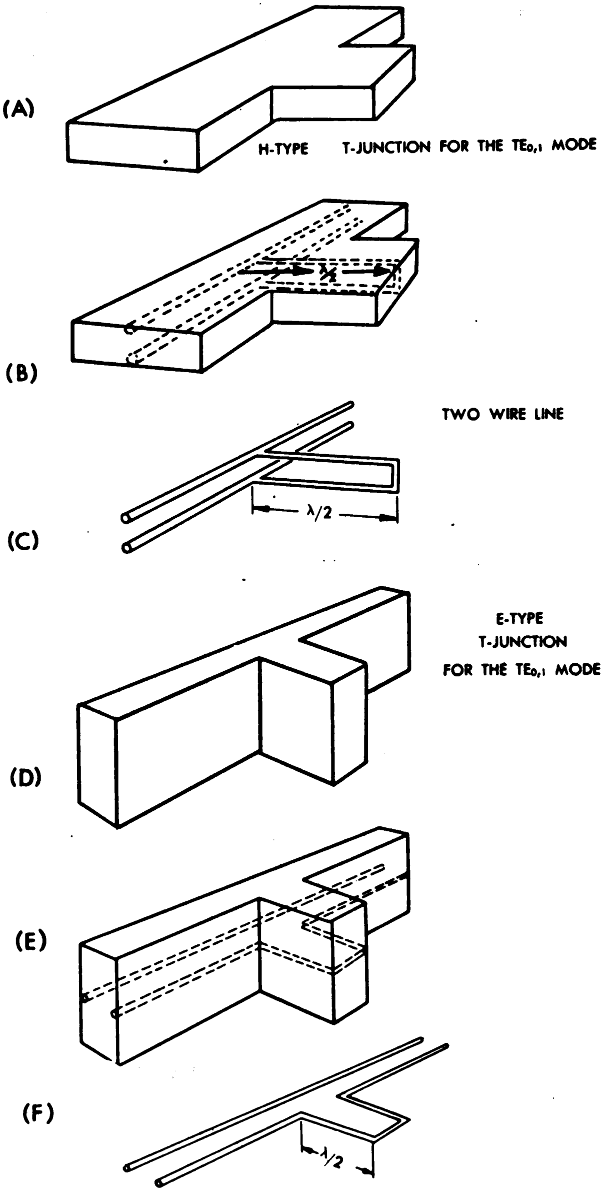


Figure 5–31.—T-junctions.

in the waveguide itself a short circuit is reflected to the center, where the E-lines are supposed to be. Since an E-line cannot exist at a short circuit, no energy will pass that point. If the shorted end of the T-section were only a quarter wave from the center of the waveguide, an open section would be reflected there and the passage of energy would be unaffected.

The E-type joint is effectively a series connection with one side of the main line ((E) of fig. 5–31). Its two-wire counterpart is shown in (F). In this case, if the section added to the waveguide is a half wave long, it will act as a short at the junction but will allow the energy to pass. The length of the added section will have to be a quarter wave in order to open the circuit at the junction and stop the ordinary flow of energy past it.

One important use of the T-joint is with the automatic transmit-receive switch (also called the TR box). In the partial transmit-receive system, illustrated in figure 5–32, an H-type T-junction is connected in the main waveguide. The switching device in this illustration is known as an anti-TR (ATR). A spark gap (TR tube) is located one quarter wave from the center of the main guide. The junction is shorted at the distance of a half wave from the center of the main waveguide. When the powerful transmitter is turned on, a spark jumps the spark gap. This causes the waveguide branch to short at that point and is inverted to an open circuit at the center. Therefore, transmitted energy can pass unhindered.

Energy received from the radar echo enters the waveguide from the other end after the transmitter is turned off. This energy is not great enough to cause a spark to jump across the spark gap. This time the shorted end of the branch reflects a short at the center of the waveguide and reflects the received energy back to another T-junction, which is located at the input to the receiver. Thus no energy is absorbed by the inoperative transmitter. There is a similar arrangement called a TR box placed in the waveguide leading to the receiver. It offers

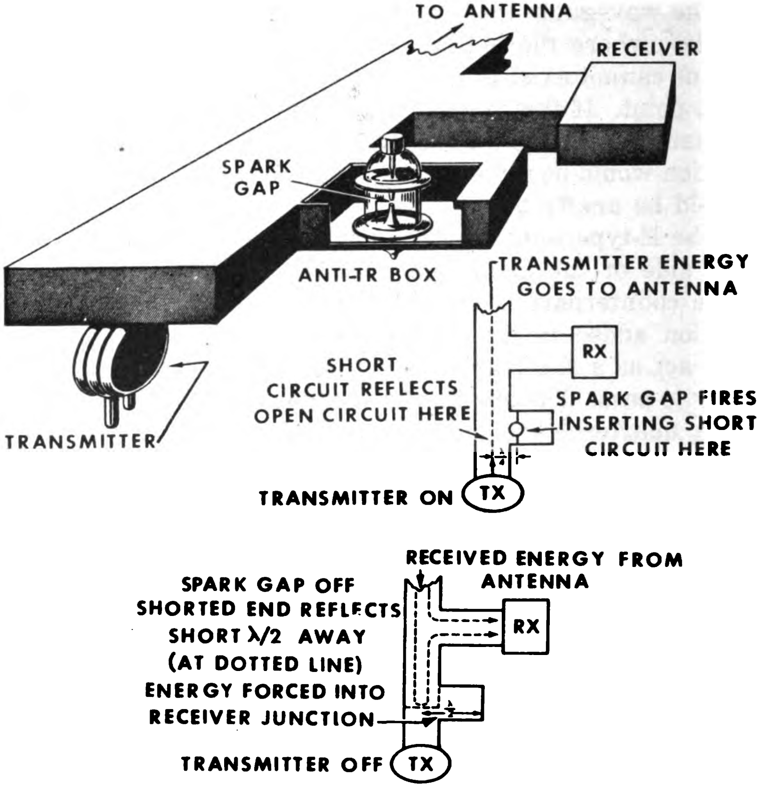


Figure 5–32.—T-junction as a transmit-receive switch.

high impedance to the transmitted pulse but low impedance to the received echo.

In practice the actual length of the sections in T-junctions is not an exact quarter or a half wavelength because the E-fields and H-fields are not perfect around the T-junction. A distortion around the junction called **FRINGE EFFECT** requires some variations from exact wavelengths. A more detailed coverage of TR tubes is given later in the chapter under the heading "Duplexer Systems."

Matching Devices

Some devices which are used in radar introduce inductance or capacitance into the circuit. Sometimes these reactances are deliberately introduced. Other times when they are present and not desired, they can be tuned out with small fins or plates in a waveguide.

Figure 5–33 shows a number of reactive plates which are deliberately used to introduce capacity or inductance in a waveguide. When these plates are employed, as shown in (A), they set up oscillations in the higher modes. Since a waveguide is too small for higher modes at the same frequency, these frequencies are not propagated, but remain in the vicinity of the plates. If the edges of the plate are vertical with respect to the plane of the H-field, the modes produced are the *TM* type. The effect of this on power flow is that of inductance across the two-wire line. This causes reflections and a shift in the standing wave pattern. The wider the space between the plates, the greater the inductive reactance.

When the partitions are arranged perpendicularly to the E-field, as in (B) of figure 5–33, a local E-field and the higher modes of oscillation are set up between the edges of the plates. These oscillations cannot be propagated but do change the dominant mode to a *TE* mode and introduce capacitive reactance. As with the *TM* mode, the wider the opening, the greater the reactance.

From these facts it would seem that a resonant circuit might be produced by combining both types of plates and leaving a small opening in a large guide, as shown in (C) of figure 5–33. This is approximately true, providing the dimensions are correct. At resonance, a resonant circuit acts like high resistance. In this condition, a small opening would introduce a high shunt resistance and the guide would in effect have connected across it a resonant circuit, since at resonance a resonant circuit acts like a high resistance.

Terminating a Waveguide

Since a waveguide is a single conductor, its characteristic impedance (Z_o) is not as easy to define as that of a coaxial line. Nevertheless, you can think of the characteristic impedance of a waveguide as being approxi-

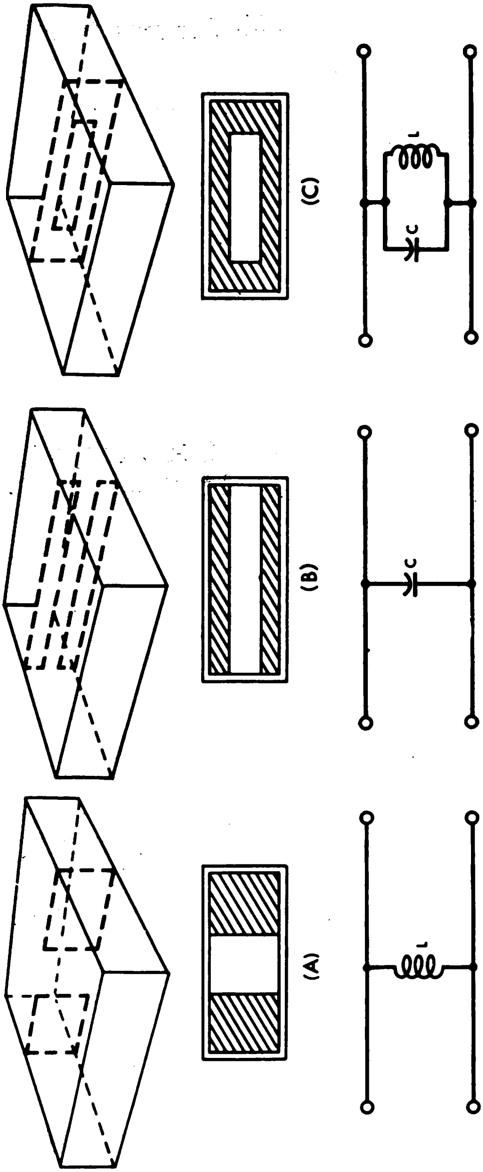


Figure 5–33.—Reactive plates in waveguide.

mately equal to the ratio of the strength of the electric field to the strength of the magnetic field for energy traveling in one direction. This ratio is equivalent to the voltage-to-current ratio in coaxial lines on which there are no standing waves.

The lowest characteristic impedance of a circular waveguide is about 350 ohms. In a rectangular waveguide, it may be any value, depending on the dimensions of the waveguide and the frequency of the electrical energy. In this guide, it is directly proportional to the narrow dimension when the other dimension and the frequency are fixed, and may vary from approximately 0 to 465 ohms.

On a waveguide there is no place to connect a fixed resistor to terminate it in its characteristic impedance as there is on a coaxial cable. However, there are a number of special arrangements which accomplish the same thing. One consists of filling the end of the waveguide with graphite-sand, as shown in figure 5–34 (A). As the fields enter the sand, currents flow in it. These currents create heat, which is instrumental in dissipating energy. None of the energy thus dissipated as heat is reflected back into the guide. Another arrangement, (B) of the figure, uses a high resistance rod, which is placed at the center of the E-field. The E-field (voltage) causes current to flow through the rod. The high resistance of the rod dissipates the energy as an I^2R loss. Still another method for terminating a waveguide is to use a wedge of high resistance as shown in (C). The plane of the wedge is placed perpendicular to the magnetic line of force. When the H-lines cut the wedge, a voltage is induced in it. The current produced by the induced voltage on flowing through the high resistance of the wedge produces an I^2R loss. This loss is dissipated in the form of heat. This permits very little energy to reach the closed end to be reflected. These terminations are used as transmitter dummy loads for testing and shop operation. They are generally constructed with external fins which give a larger radiation surface, thus better heat dissipation.

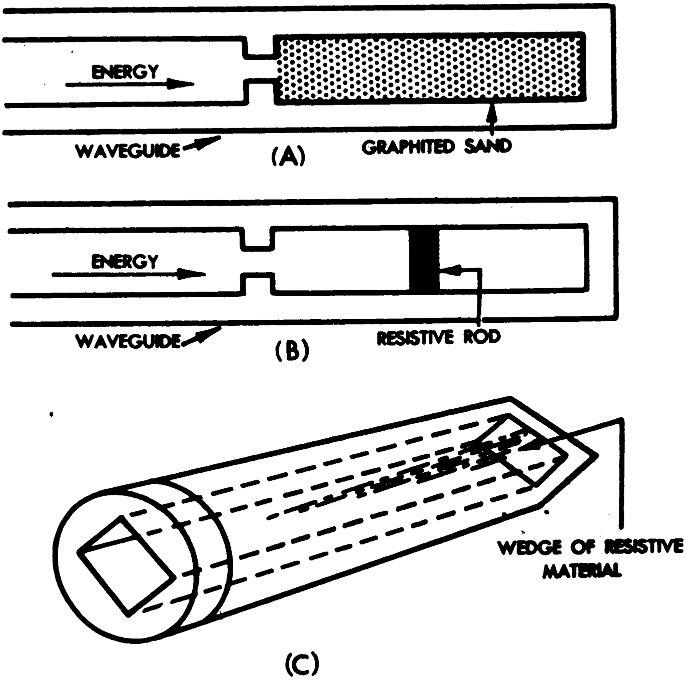
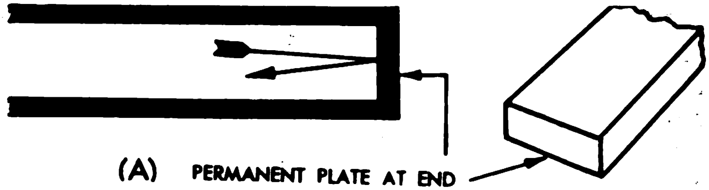


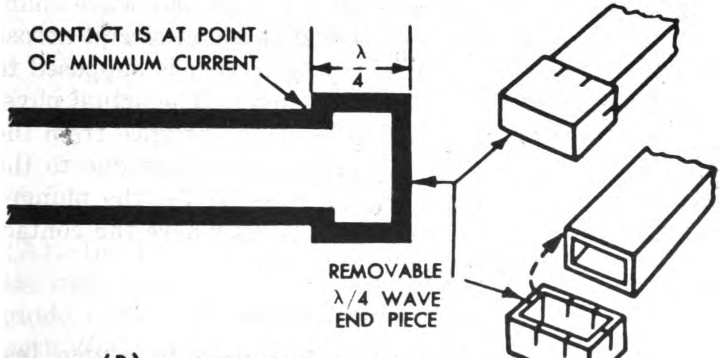
Figure 5–34.—Termination for minimum reflections.

Each of the preceding terminations is designed to match the impedance of the guide in order to insure a minimum of reflection. On the other hand, there are many instances where it is desirable for all the energy to be reflected from the end of the waveguide. The best way to accomplish this is to weld a permanent metal plate at the end of the waveguide. (See fig. 5–35 (A).)

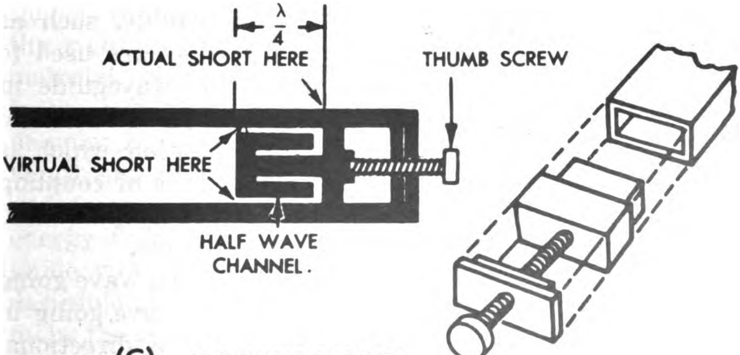
When it is necessary that the end be removable, a removable end plate is attached to the end of the guide. For this method to be satisfactory, the contact between the guide and the end plate must be exceptionally good in order that the H-field will not be attenuated when current flows. Perfect contact is not required when the connection is made at a point of minimum current. This point is located a quarter wave from the end. When connection is made at this point, a cup is used, as in (B),



(A) PERMANENT PLATE AT END


CONTACT IS AT POINT OF MINIMUM CURRENT
λ/4
REMOVABLE λ/4 WAVE END PIECE

(B) REMOVABLE SHORT CIRCUIT



(C) ADJUSTABLE PLUNGER

Figure 5–35.—Termination for maximum reflection.

instead of the end plate. This cup is a quarter wave long and large enough to fit over the end of the guide. The voltage between opposite sides of the cup opening is high, but as the reflected H-field cancels the incident H-field, the resulting current is very small and reflection is at a minimum.

When the end must be adjustable, the contact must be nearly perfect. However, it is impossible to get a perfect contact. The best arrangement, one which is similar to the choke joint previously explained, consists principally of an adjustable plunger which fits into the guide as shown in (C). The walls of the waveguide and the plunger form a half-wave channel. The half-wave channel is closed at the end and reflects a short circuit across the other end, where a perfect connection is supposed to exist between the wall and the plunger. The actual physical contact is made at a quarter-wave distance from the short circuit, where the current is minimum due to the standing waves. This makes it possible for the plunger to slide loosely in the guide at a point where the contact resistance to current flow is very low.

Directional Couplers

Directional couplers are devices used to couple test equipment to a waveguide and are commonly included as a permanent section of the transmission line. They are used for measuring power in the waveguide, such as radar transmitter power output; also, they are used to inject a signal of known power into the waveguide in order to determine receiver sensitivity. Directional couplers have a fixed amount of attenuation which is generally stamped on the coupler. This type of coupling is used extensively for it contains no moving parts and injects negligible reflection into the waveguide.

A directional coupler couples energy from a wave going in one direction in a guide but not from a wave going in the other direction. There are two types of directional couplers—TWO-HOLE and ONE-HOLE.

A two-hole directional coupler is shown in figure 5–36

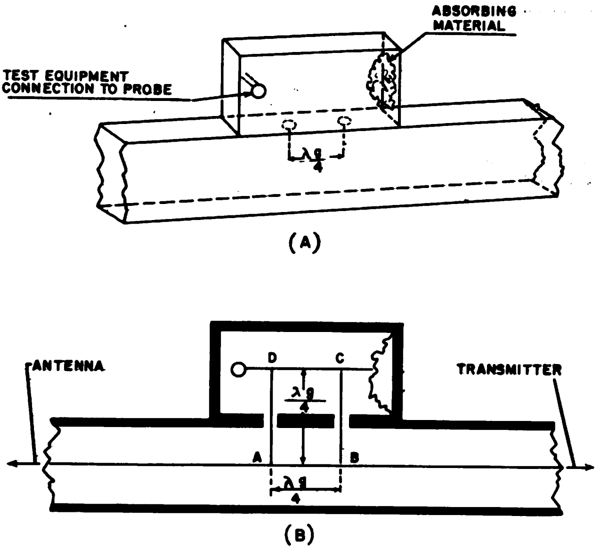


Figure 5–36.—Two-hole directional coupler.

(A). In (B) of this figure note that a wave traveling to the right in the bottom guide has two paths into the upper guide, one from *A* to *D* directly and one from *A* to *D* via path *B–C*. However, the two paths differ in length by one-half wavelength, so that waves arriving by both paths interfere destructively, and no wave proceeds to the left in the upper guide. Also note that waves going to the right in the upper guide are lost in the absorbing material. However, if the wave in the main guide moves to the left, then the two paths by which waves arrive at point *D* (path *B–C–D* and path *B–A–D*) are equal in length and a wave proceeds to the left in the upper guide, and is picked up by the test equipment probe. Thus, energy from the transmitter passing to the left in the guide will reach the test equipment connection with a desirable amount of fixed attenuation, but energy picked up by the antenna which passes to the right will not reach the test equipment. A signal injected at the test equipment probe will travel toward the receiver (which is be-

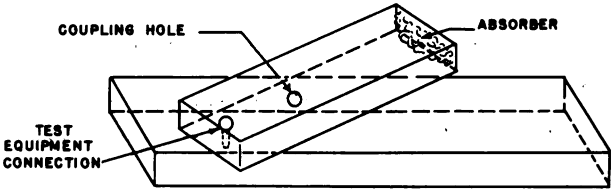


Figure 5–37.—One-hole directional coupler.

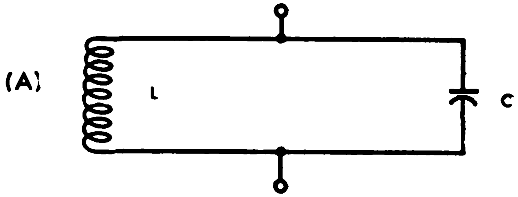
tween the directional coupler and transmitter) but will not pass to the left toward the antenna.

The one-hole directional coupler is shown in figure 5–37. The coupling from the main guide to the branch is done by the electric and magnetic fields. The phase of the magnetic field is changed by the coupling so that the waves in the branch are cancelled in one direction and reinforced in the other. In contrast to the two-hole coupler, the one-hole coupler transmits a wave in a direction opposite to that of the wave in the main line. The proper coupling to secure the directional characteristic depends on the orientation angle between the main guide and the branch guide and the size of the hole between the two guides.

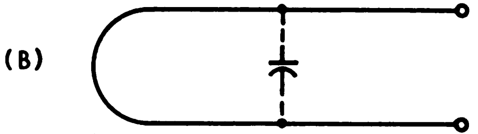
The branch guide is terminated at one end with a resistive material to absorb unwanted power, and the other end is connected to a probe for connection to the test equipment.

CAVITY RESONATORS

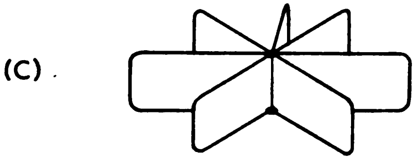
In ordinary radio work the conventional low-frequency resonant circuit consists of a coil and capacitor which are connected either in series or in parallel as in figure 5–38 (A). To increase the resonant frequency, it is necessary to decrease either the capacitance or the inductance or both. However, a frequency is reached where the inductance is a half-turn coil and where the capacitance consists only of the stray capacitance in the coil. At extremely high frequencies this resonant circuit would consist of a coil about an inch long and a quarter inch across.



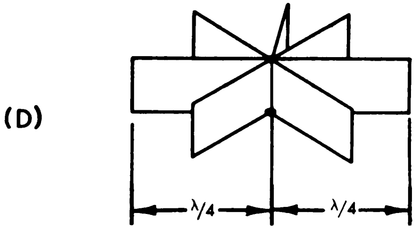
CONVENTIONAL LOW FREQUENCY CIRCUIT



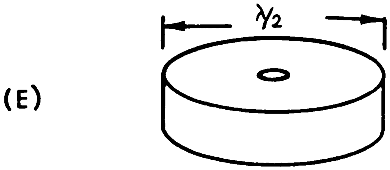
COIL-CONDENSER COMBINATION REDUCED FOR HIGHEST RESONANT FREQUENCY



HALF TURN LOOP IN PARALLEL



QUARTER WAVE SECTIONS



CLOSED METAL CONTAINER

Figure 5–38.—Development of cavity from quarter-wave sections.

In this circuit the current handling capacity and breakdown voltage for the spacing would be low.

The current carrying ability of a resonant circuit may be increased by adding half-turn loops in parallel. This does not change the resonant frequency appreciably because it adds capacitance in parallel, which lowers the frequency, and inductance in parallel, which increases the frequency. As the effects of each cancel, the frequency remains about the same.

In figure 5–38 (C), several half-turn loops are added in parallel. In (D) are shown several quarter wave Lecher lines, in parallel, which are resonant when they are near a quarter wavelength. When more and more loops are added in parallel, the assembly eventually becomes a closed resonant box as shown in (E) which is a quarter wave in radius or, in other words, a half wave in diameter. This box is called a RESONANT CAVITY.

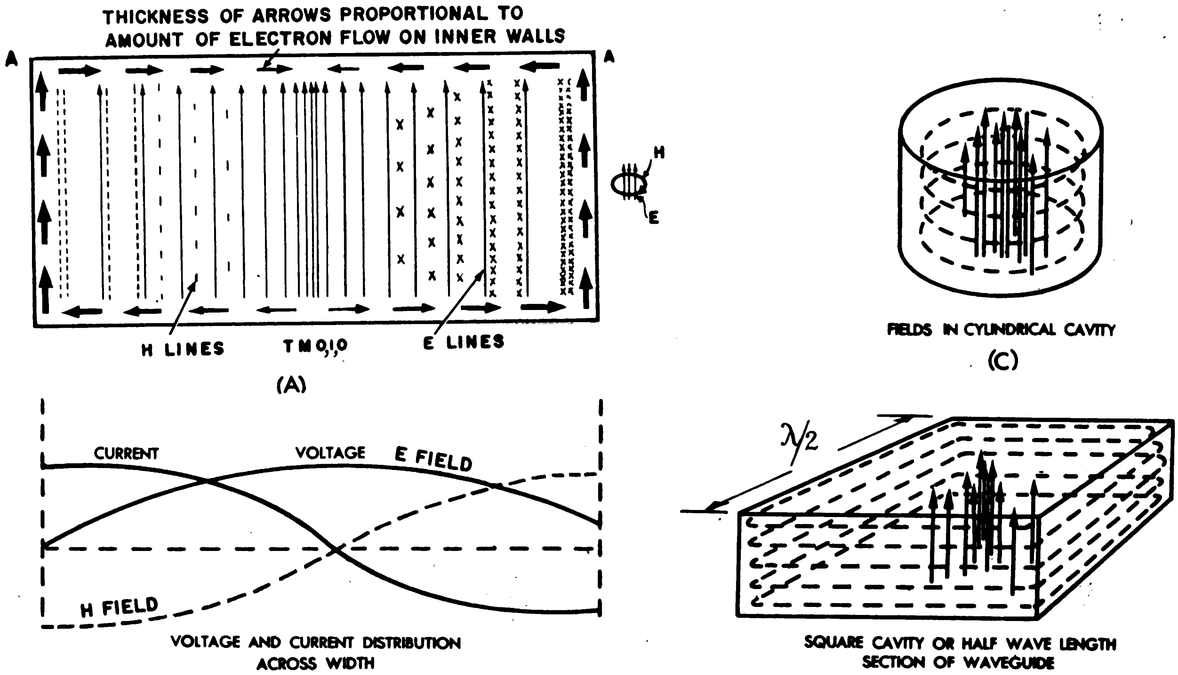
A resonant cavity displays the same resonant characteristics as a tuned circuit composed of a coil and capacitor. In it there are a large number of current paths. This means that the resistance of the box to current flow is very low and that the *Q* of the resonant circuit is very high. While it is difficult to attain a *Q* of several hundred in a coil of wire, it is fairly easy to construct a resonant cavity with a *Q* of many thousand. Although a cavity is as efficient at low frequencies as at high frequencies, the large size required at low frequencies prohibts its use at those frequencies. For example, at one megacycle a resonant cavity would be a cylinder about 500 feet in diameter. When the frequency is in the vicinity of 10,000 megacycles, the diameter of the cavity is only 0.6 inch. This makes the cavity smaller than a conventional tuned circuit. Therefore, equipment which operates at a frequency of approximately 3,000 mc. or above employes resonant cavities extensively.

The Fields in a Cavity

A resonant cavity may be compared to a waveguide, since its operation is best described in terms of the fields rather than in terms of the currents and voltages present. As in waveguides, the different field configurations in cavities are called MODES. In figure 5–39, in which the dominant mode of the cylindrical cavity is shown, note that in (A) the voltage is represented by E-lines between the top and the bottom of the cavity. The current, due to skin effect, flows in a thin layer of the surface of the cavity. The strength of the current is indicated by arrows of various lengths. The magnetic field is strong where the current is high. The strongest H-field is at the vertical walls of the cylinder and diminishes toward the center where the current is zero. This is due to the standing waves on the quarter-wave section. The E-field is maximum at the center and decreases to zero at the edge, where the vertical wall is a short circuit to the voltage. The curves of E-field and H-field density are shown in figure 5–39 (B); two types of cavities with their fields represented are shown in (C) and (D).

The modes in a cavity are identified by the same numbering system that is used with waveguides, except that a third subscript is used to indicate the number of patterns of the transverse field along the axis of the cavity (perpendicular to the transverse field). For example, the cylindrical cavity shown in (C) of figure 5–39 is a form of circular waveguide. The axis is the center of the circle. The transverse field is the magnetic field. Therefore, it is *TM* (transverse magnetic). Around the circumference there is a constant magnetic field. (The H-lines are parallel to the circumference.) Therefore, the first subscript is zero. The distance across the diameter is one-half wave. Thus the second subscript is one. Through the center, along the axis, the H-field strength is a constant zero. This makes the third subscript zero. Therefore, the complete description of the mode is $TM_{0,1,0}$.

When a section of waveguide which is one-half wave long is closed on both ends in the form of a rectangular cavity, as shown at (D) in figure 5–39, standing waves are set up and resonance occurs. The simple mode in this cavity is the same as the dominant mode of a


THICKNESS OF ARROWS PROPORTIONAL TO AMOUNT OF ELECTRON FLOW ON INNER WALLS
A
A
H
E
H LINES
T M 0,1,0
E LINES
CURRENT
VOLTAGE
E FIELD
H FIELD
λ/2

(A)

FIELDS IN CYLINDRICAL CAVITY

(C)

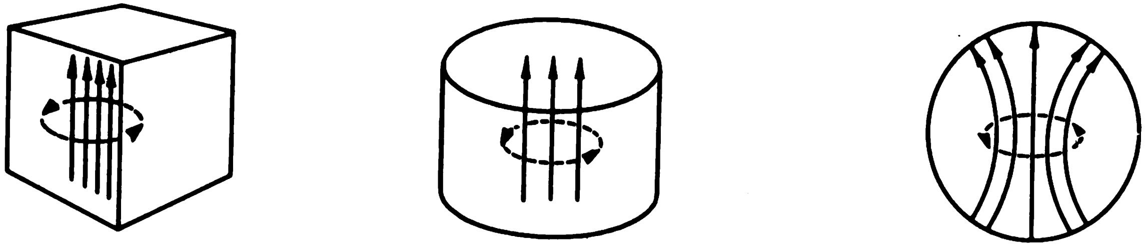
VOLTAGE AND CURRENT DISTRIBUTION ACROSS WIDTH

(B)

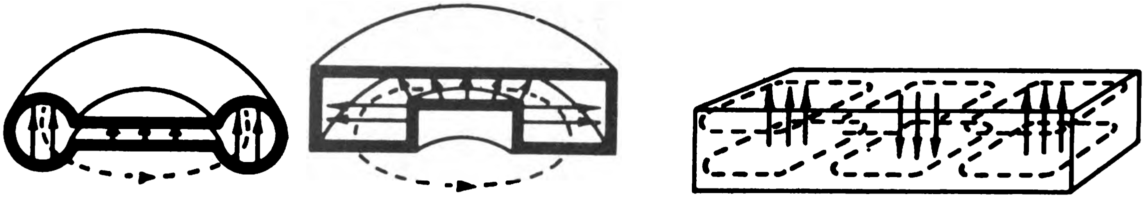
SQUARE CAVITY OR HALF WAVE LENGTH SECTION OF WAVEGUIDE

(D)

Figure 5–39.—Voltages, currents, and associated fields for simple modes in a cavity resonator.



CUBE (Q = 28,000) CYLINDER (Q = 31,000) SPHERE (Q = 26,000)



DOUGHNUT-SHAPED CYLINDRICAL RING (Q = 26,000) SECTION OF WAVEGUIDE

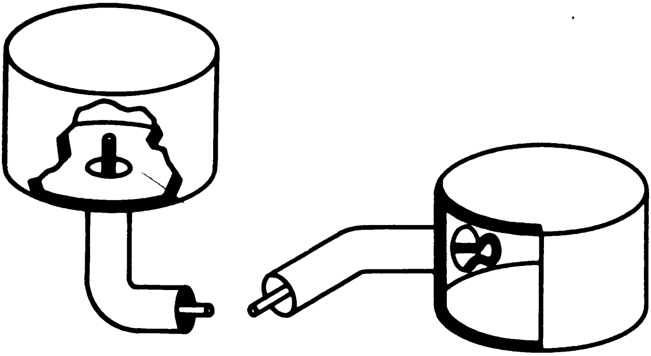
Figure 5—40.—Several types of cavities.

rectangular waveguide, that is, it is $TE_{0,1}$. The third subscript of the mode, which is determined by the plane of the E-field, is one. Thus the complete description of the simple mode in the rectangular cavity in (D) is $TE_{0,1,1}$ (transverse electric).

Cavities may have various physical shapes, for any chamber enclosed in conducting walls resonates at several frequencies and produces a number of modes. Figure 5–40 shows several types of cavities. (Note the large approximate Q values indicated in some cases.) Of those shown the cylinder type cavity is useful in wavemeters or in frequency measuring devices. The cylindrical ring type is used in superhigh-frequency oscillators as the frequency determining element. The section of waveguide which is shown diagrammatically is used in some radar systems as a mixing chamber for combining signals from two sources.

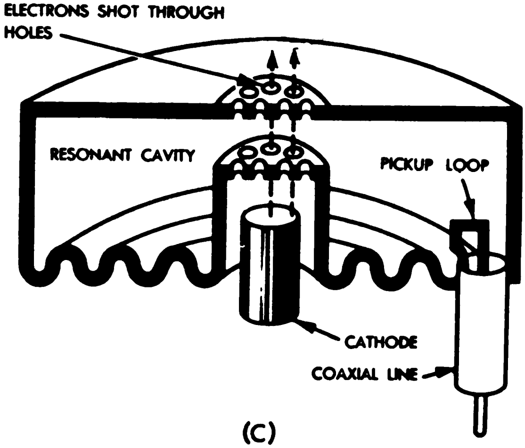
Exciting the Cavity

There are three principal ways in which energy can be inserted into and removed from a cavity resonator (fig. 5–41). The first is by inserting a probe into the cavity. The current which flows in the probe sets up E-lines parallel to it, and they in turn start oscillation. Another method uses a magnetic loop. The loop is placed in the region where the magnetic field will be located. The currents in the loop start an H-field in the cavity. Either of these methods can be used to remove energy from or to put energy into the cavity. A third method uses a cylindrical ring type cavity. In this method the energy is placed into the cavity by clouds of electrons, which are virtually shot through the holes in the center of a perforated plate. As each cloud goes through, it creates a disturbance in the space inside the cavity until a field is set up. In terms of current, it may be said that the approaching cloud of electrons makes the perforated plate positive by repelling the electrons away from it. This current starts an H-field. Energy may be removed from the cavity by placing a loop at the outside edge.



(A)
PROBE COUPLING
TM$_{0,1,0,}$ MODE

(B)
LOOP COUPLING TM$_{0,1,0,}$ MODE



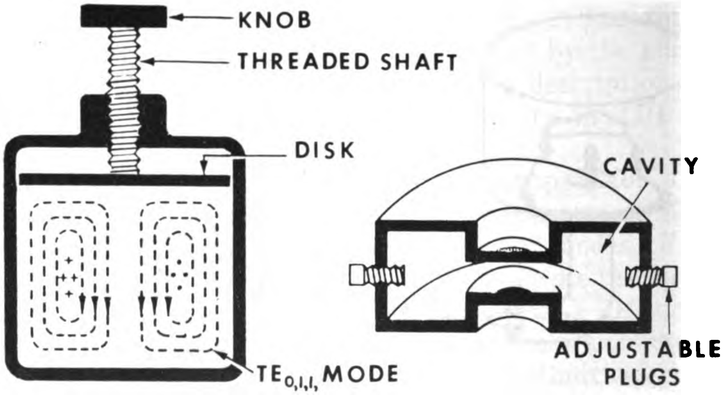
(C)
CYLINDRICAL RING

Figure 5–41.—Methods of exciting the cavity.

Other methods of feeding the cavity are discussed later in the analysis of systems using cavities.

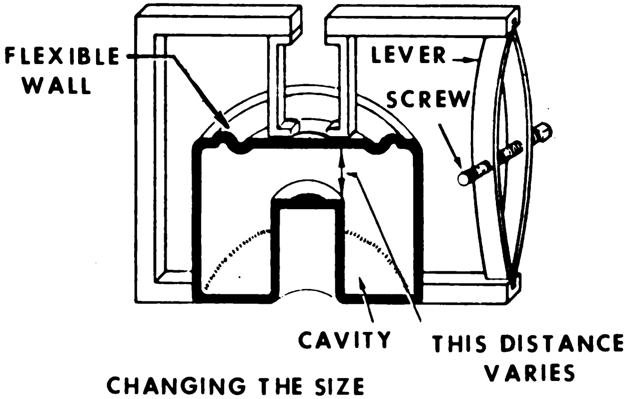
Varying the Resonant Frequency of the Cavity

Three methods for setting the resonant frequency in a cavity are shown in figure 5–42. One method uses a cylindrical cavity with an adjustable disk. When a $TE_{0,1,1}$



CYLINDRICAL CAVITY WITH ADJUSTABLE DISK

CAVITY WITH ADJUSTABLE PLUGS



CHANGING THE SIZE OF CAVITY

Figure 5–42.—Methods of changing the frequency of a cavity.

mode is used, the size of the cylinder may be changed along the axis to change the resonant frequency. The smaller the volume of the cavity, the higher the resonant frequency. The movement of the disk may be calibrated in terms of frequency. Usually in the case of high-frequency equipment a micrometer scale is used to indicate the position of the disk, and a calibration chart is used to determine the frequency.

A second method for varying the frequency employs threaded plugs which are inserted in the side of the cavity. The plug reduces the strength of the magnetic field in the cavity in a manner similar to reducing the inductance of the tuned circuit. The deeper the plug extends into the cavity, the higher the frequency.

In a third method, the interior of the cavity, which is part of the interior of a vacuum tube, is sealed and evacuated. In this method a frequency change occurs whenever the top and bottom of the cavity are moved away from (or toward) each other. This is accomplished by turning a screw at the end of a lever. As the distance varies, the volume and the capacity between the top and bottom of the cavity are changed. As the change in capacity is the chief result of the change in the distance between the plates, the resonant frequency is inversely proportional to the distance from the top to the bottom.

Another way of changing the frequency is by changing the method of exciting the circuit. This can be done by tuning the exciting loop either capacitively or inductively. This can occur either accidentally due to improperly tuned circuits or deliberately as a means of tuning the cavity.

Uses of Cavities

There are many ways in which resonant cavities are used in electronics. For example, the klystron in figure 5–43 (A), which is usually the local oscillator in a radar receiver, is tuned by the resonant cavity previously shown in figure 5–41 (C). In the klystron, the cathode emits electrons which pass through the grids toward the plate. When they pass through this area, they disturb the field and excite the cavity. The resonant frequency is changed by the spacing of the grids. Another use of resonant cavities is the series of cavities arranged around a circle in the magnetron oscillator as shown in (B) of figure 5–43. These cavities, which are coupled one to another through capacitance at the openings, are excited by moving electrons. The energy from the electrons is

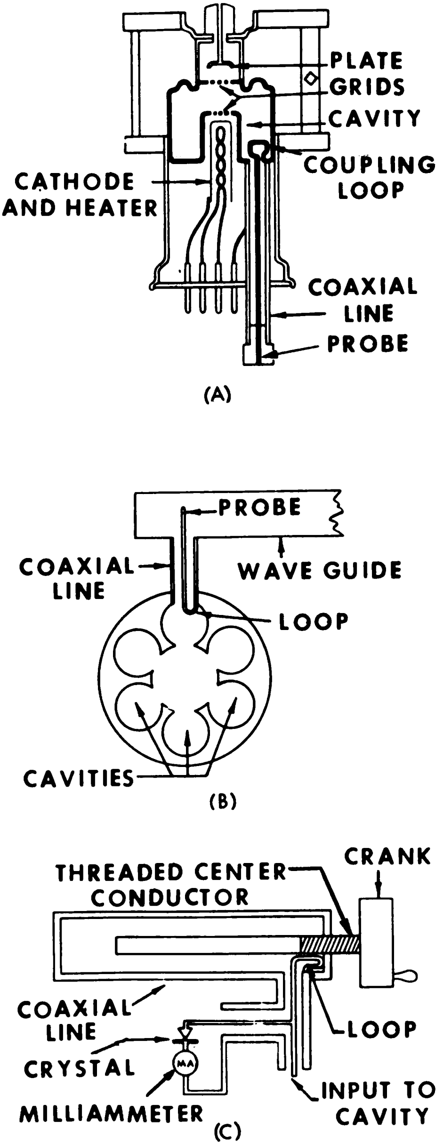


Figure 5—43.—Uses of cavities.

passed around the ring from all cavities to the one with a loop. This one serves to transfer the energy to the waveguide. From the waveguide, the energy travels to the antenna where it is radiated into space.

A tunable magnetron is used in some equipments in order that the transmitter frequency may be varied. Although most magnetrons used are of the fixed frequency type, they possess characteristics, other than cavity dimensions, which will determine their output frequency and mode of oscillation. These characteristics will be explained in chapter 6.

Cavities also may be used as wavemeters (devices for measuring frequency). One type of wavemeter consists of a CYLINDRICAL cavity with an adjustable disk. Another type operates essentially like this but employs a COAXIAL cavity and operates in a different mode, as shown in figure 5–43 (C). In the second type, the signal is introduced into the cavity by a loop which is located in one end of the cavity. The signal is strongest when the loop is at a high current point in the standing wave. The standing wave of current is maximum at odd multiples of quarter-wave distances from the open end of the guide. When the length of the threaded center conductor is varied by the crank, the distance from the open end to the loop can be made equal to an odd multiple of quarter waves. This causes the current maximum to occur at the loop and this in turn causes the input impedance at the loop to be zero. When the distance from the open end of the guide to the short (closed end) is a quarter wave or any odd multiple of a quarter wave, the current introduced into the guide will be at the location of the parallel resonance.

The threaded shaft shown in (C) of figure 5–43 can be calibrated either in wavelength or in frequency. The input impedance can be indicated by any suitable indicator, such as a crystal rectifier and a d-c milliammeter. In operation, when the input impedance is zero, the signal is shorted out. In this case, the meter will read zero.

To check the frequency for a zero impedance reading, turn the crank until a dip occurs on the meter and read the frequency on the threaded shaft.

A cavity can be used as a mixing chamber. When two or more signals are put into a cavity, it is possible to remove the combination of their signal voltages. An example of this use is discussed later in this chapter.

Another use of a cavity is as an impedance matching device such as shown in figure 5–44, which illustrates a small waveguide connected to a large waveguide. Normally, when you connect the two waveguides together, there will be reflections in the waveguides. However, a small section of waveguide or an intermediate size can be matched without reflections, provided that plates are used at the junction to cancel the reactive effects which result from the different sizes of the waveguides. Thus, although the cavity itself has standing waves in it, there are none on either waveguide.

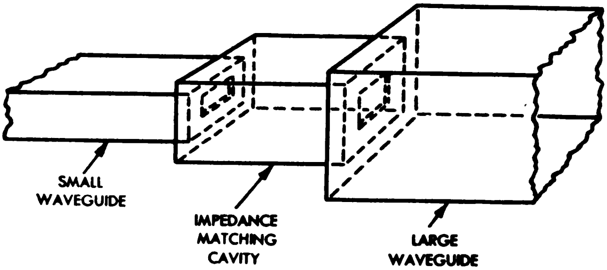


Figure 5–44.—Cavity used for matching an impedance device to one of different impedance.

The ECHO BOX uses a cavity as a RINGING CIRCUIT in a resonant circuit. The ringing circuit gets its name from the fact that it oscillates for many cycles after being started by an external circuit. In this use the cavity is resonant, and once oscillation is started it will continue for a few microseconds after the source of voltage has been removed. This action is analogous to a bell which will ring for many seconds after being struck. In the 3 cm. (10,000 mc.) ringing circuit shown in figure 5–45,

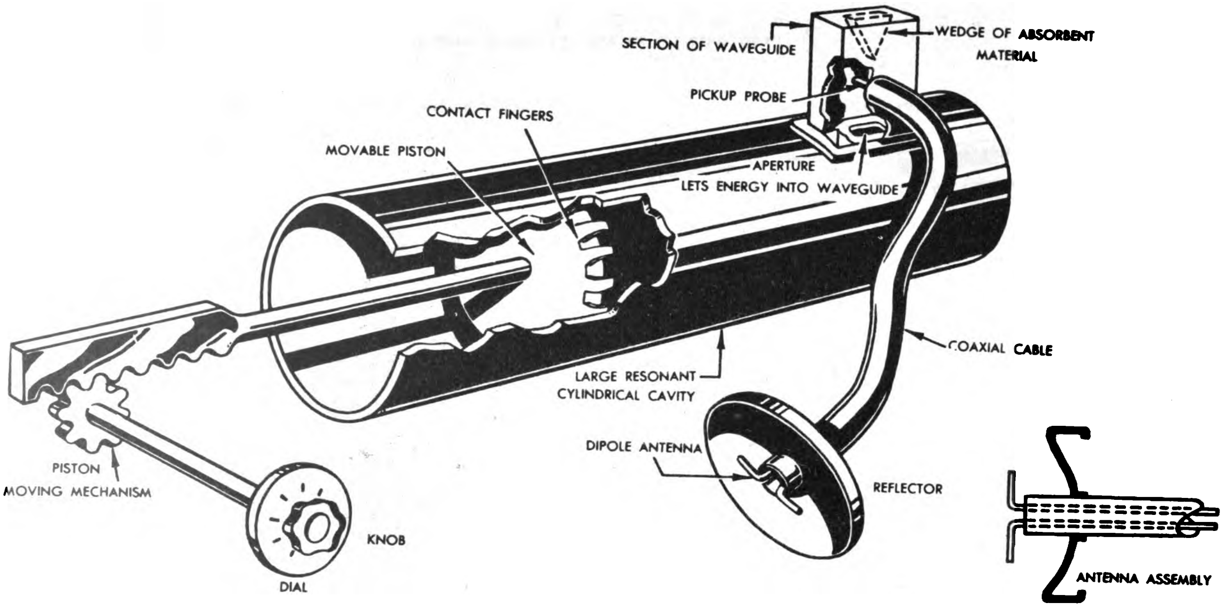


Figure 5–45.—A 3 cm. ringing circuit.

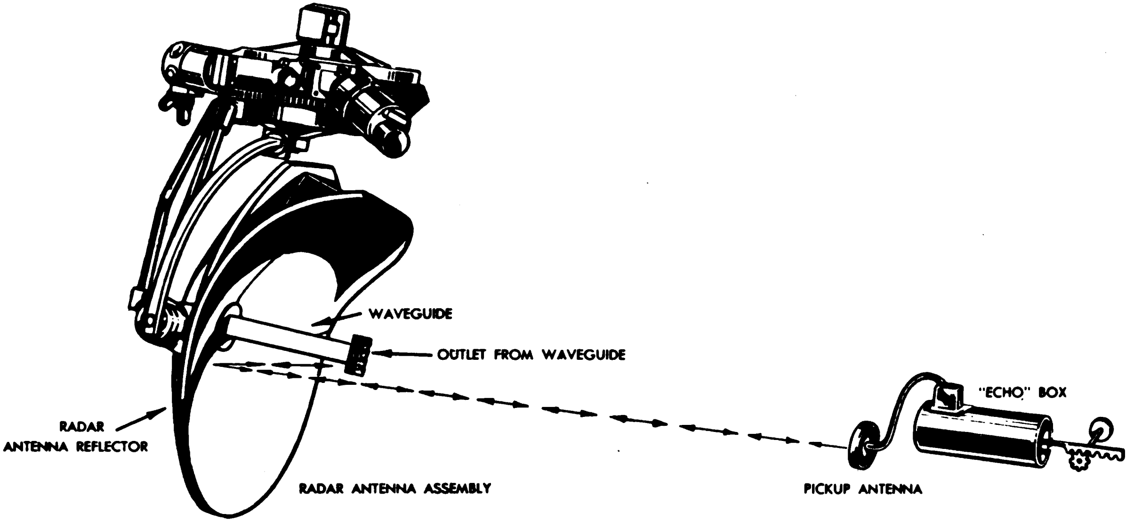


Figure 5–46.—Cavity as an echo box.

the cavity itself is a cylinder which is three inches in diameter and 10 inches long. A movable piston in the cylinder determines the volume of the cavity. As the position of the piston is calibrated in megacycles, it serves as a frequency meter. The cavity is energized by a signal which enters through a small aperture from the small section of waveguide. The energy is induced in the short waveguide by a probe, which in turn is fed by a coaxial cable. A good method of feeding the signal into the coaxial cable is by connecting a small antenna to the end of the cable.

The resonant cavity just described is used as an echo box in figure 5–46. Echo boxes are resonant cavities which have a very high *Q*. They are used with microwave radar sets to provide an artificial or "phantom" target, which may be used to tune the receiver to the transmitter when no real targets are available.

This is how the echo box works. When the radar transmitter pulses, energy is fed into the echo box cavity. There it sets up violent oscillations, and the cavity itself acts like a transmitter for a period of about 20 to 40 microseconds after the pulse generated by the radar transmitter ends. (When you view the oscillations on an oscilloscope, their waveshapes appear as shown

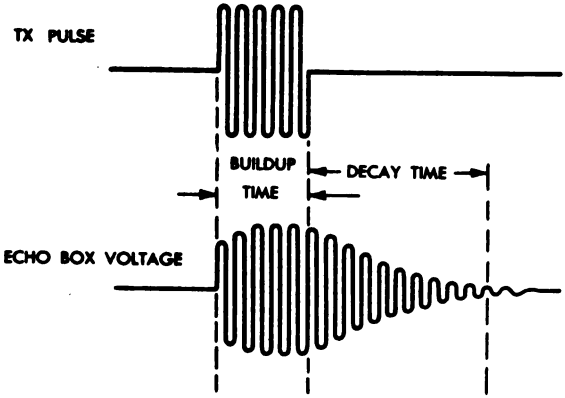


Figure 5–47.—Oscillations in echo box.

in figure 5–47.) The energy which the echo box radiates is picked up by the small pickup antenna located in front of the radar antenna assembly and fed into the radar receiver. The receiver displays it on the indicator screen. Since this signal simulates a radar echo, the cavity which transmitted it is appropriately called an ECHO BOX. It is thus a means of supplying a signal for tuning the receiver when there is no real target available. The echo box may also be connected directly into the waveguide's directional coupler.

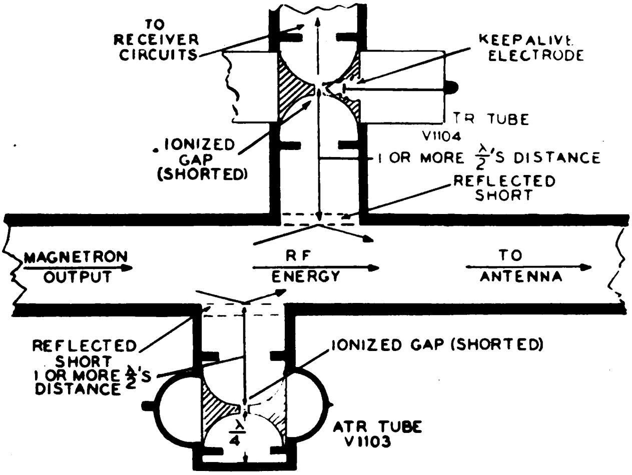
DUPLEXER SYSTEMS

Basic Systems

In present-day aviation radar it is desirable to use a single antenna for both transmitting and receiving. A fast acting electronic switching device is required to disconnect the receiver from the transmission line during periods of transmission and to disconnect the magnetron during periods of reception. The receiver must be disconnected in order to protect the receiver input section and crystal since the magnetron output may be several megawatts. This function is performed by the TR tube. The magnetron must be disconnected during periods of reception in order to allow all available reflected energy from the target to pass to the receiver. This function is performed by the ATR tube.

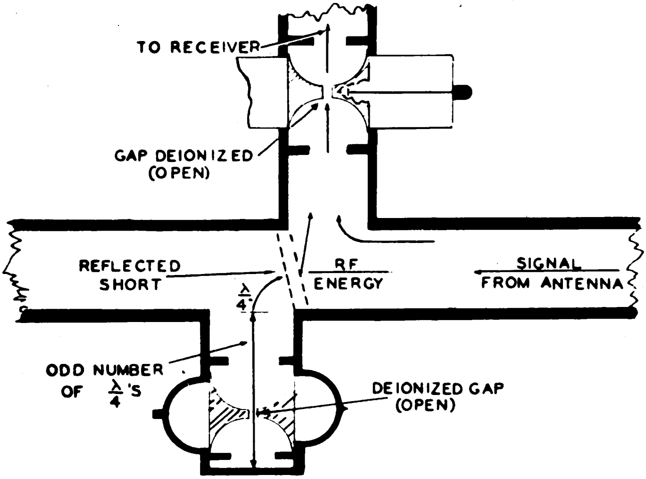
The device used to perform this switching is known as a DUPLEXER. In performing this action the duplexer makes use of one or more gas-filled tubes known as transmit-receive (TR) tubes. These tubes are a spark gap enclosed in a gas-filled envelope. They are practical for this function because of the tremendous power difference between the transmitted and received signal. Ionization occurs during the transmitter pulse period and causes the tube to function as a short circuit or closed switch. During the reception period these tubes remain de-ionized because the received signals are low power; thus they appear as an open circuit.

A simplified cross-section view of a waveguide system



TRANSMISSION OF PULSE

(*A*)



SIGNAL RECEPTION

(*B*)

Figure 5–48.—Duplexer action in a waveguide.

duplexer is shown in figure 5–48. The TR tube (*B*1104) is used to protect the crystal mixer and the receiver circuits during transmission. When the magnetron fires, sufficient voltage is impressed on the TR tube to ionize it completely and break down the gap, producing an effective short circuit. Since the TR tube is placed a number of half wavelengths from the waveguide, a short circuit existing within the tube reflects an effective short circuit at the entrance of the receiver T-junction. This prevents the RF pulse from reaching the crystal and the receiver, but permits it to flow freely to the antenna. The operation of the duplexer during transmission is shown in figure 5–48 (A). After the pulse has passed, the tube de-ionizes and again presents a high impedance at the entrance of the receiver T-junction. This allows any reflected signals to reach the crystal mixer. Duplexer action during signal reception is shown in (B) of the same figure.

The ATR tube (*V*1103) is located a number of half wavelengths from the waveguide; thus when the transmitter fires, which breaks down the gap of the ATR tube, an effective short circuit is reflected across the opening at this T-junction. This causes the wall to appear continuous to the transmitted energy; thus the energy passes unimpeded. The ATR waveguide section is so constructed that an odd number of quarter wavelengths exists between its shorted end and the opening to the waveguide. During signal reception, the ATR tube is de-ionized and causes an effective open circuit to appear at the opening of the ATR junction. This causes a reflected short across the waveguide adjacent to the TR opening, as illustrated in figure 5–48 (B). The received signal is guided into the TR cavity instead of continuing to the magnetron.

Notice that both TR's in this illustration are in the wide section of the waveguide, thus they are an E-type junction whereas the application of an ATR, previously explained in this chapter under the heading "T-junction," was an H-type.

Characteristics of TR Tubes

At atmospheric pressure approximately 30,000 volts are required to break down a spark gap of one inch. Approximately 50 volts is required to maintain ionization; this is known as running-voltage. Once the voltage has been removed a lapse of approximately 10μsec. is required for de-ionization. By enclosing the spark gap in a partially evacuated envelope, the breakdown voltage will be reduced as well as the running-voltage and the de-ionization time.

Figure 5–49 shows the voltage envelope that appears across the TR spark gap and the receiver input when the transmitter fires. For efficient transmitter operation it is important that the power dissipated by the TR tubes is small as compared to transmitter power. Also the TR tubes must maintain a properly matched line while they are fired. During the time that the transmitter is fired the TR tubes must afford sufficient attenuation to prevent the power that leaks into the receiver from being of so great an amplitude that it may damage the receiver crystal. To allow scope presentation of nearby targets it is necessary that the TR tubes de-ionize rapidly.

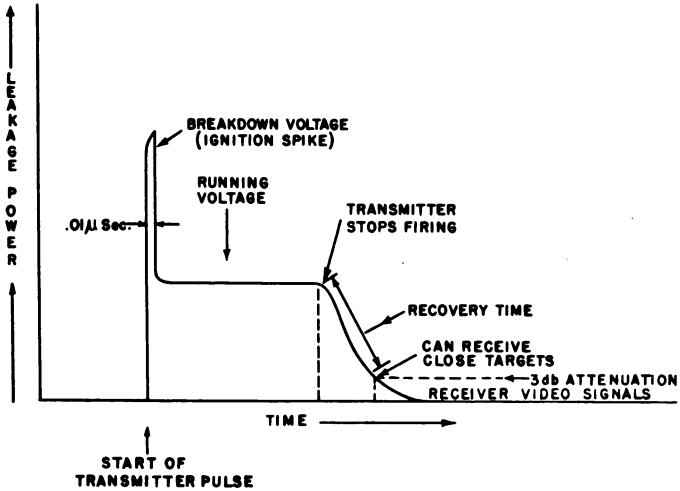


Figure 5–49.—Voltage envelope seen across the TR spark gap when the transmitter is fired.

The amplitude and duration of the ignition spike must be reduced to the extent that the receiver crystal will not be damaged. This is accomplished by providing a keep-alive voltage to the TR tubes. TR tubes that perform this function contain three electrodes, two of which provide the main gap which does the switching, and the third, called the keep-alive electrode. This keep-alive electrode is mounted close to one of the other electrodes. It is connected to a negative potential of approximately 800 volts which causes ionization of the gas between it and the main electrode. The ions insure almost instantaneous firing of the TR tube for each pulse of the magnetron.

A CRYSTAL SHUTTER is frequently used as a protective device for the receiver crystal while the radar is turned off. When the radar is off no keep-alive voltage is applied to the TR tube; thus there is the possibility that energy radiated from nearby radars may damage the crystal. A typical crystal shutter assembly may consist of a waveguide gate and a solenoid that controls the position of the gate. When the radar system is not in operation the solenoid is not energized and the position of the gate is such that it causes a virtual short circuit across the waveguide path to the receiver input. When the radar system is in operation, the solenoid is energized and causes the gate to move from the waveguide to a position where it will offer no interference to signals traveling to the receiver.

Protection of the receiver may be improved by installing a transformer between the waveguide and the TR tube. As a result of the step-up ratio of the transformer the voltage across the TR tube will be greater than that across the line. This increase in voltage causes the tube to ionize faster, thus more receiver protection. An added benefit of the transformer action is less transmitter power loss since there is less voltage drop across the line at the TR T-junction. Resonant transformers are used for this purpose in microwave TR systems.

The cavity of the TR tube is designed to have a high *Q;* this necessitates careful tuning. Tuning is accomplished by means of a tuning screw which varies the separation of the main electrodes. Some equipments use a tunable ATR tube which must also be tuned to the transmitter. A detuned TR will cause receiver sensitivity to drop. A means of determining if the TR is detuned is to compare the sensitivity of the receiver at the image frequency with the sensitivity of the receiver at the fundamental frequency. Before this comparison can be made it is necessary to know what this difference in sensitivity should be. This data may be obtained by making tests on a radar that is functioning in accordance with its specifications. The data obtained may be used for comparison reading.

RECOVERY TIME.—Recovery time may be described as the time required for a TR tube to de-ionize effectively. Specifically it is the time between the end of the transmitted pulse and the time where the received signal is attenuated 3 db (half-power point). This recovery time varies from approximately 6 μsec. to 36 μsec., depending upon the type of gas used in the TR tube as well as the amount of transmitter power. The length of the recovery time period is determined by the length of time required for the free electrons and ions, that remain in the gas after the transmitter pulse has been removed, to recombine. This recovery period will vary depending upon the type of gas used in the TR tube.

LIFE OF TR TUBES.—The life of a TR tube is relatively short as compared with most other tubes. A major factor that shortens the life of these tubes is the gradual loss of gas pressure. This is the result of the gas being absorbed by tube electrodes. Another reason for the short life of the tubes is that molecules of gas become embedded in the electrodes; this is caused by glow and keep-alive discharge. One means of lengthening the life of these tubes is to maintain the keep-alive current within designed limits. Some equipments provide test points for making current checks. A reading lower than the normal minimum current will most likely be an indication that the TR tube is not operating at proper efficiency.

The most general indications of defective TR tubes are as follows:

1. Receiver recovery time will increase.
2. Receiver sensitivity will decrease.
3. Receiver crystals will become defective. (This may be detected by a rapid increase in crystal back current.)

Some radars use the same tube for TR and ATR operation. In some cases a tube that will not operate as a TR may be used satisfactorily as an ATR; thus for emergency operation the tubes may be interchanged.

Microwave Coaxial Line Duplexer Systems

At frequencies up to approximately 3,000 megacycles coaxial lines are generally used for RF transmission. The reason for using this type line is that at these frequencies should a waveguide be used the size would be so great that it would be impractical. Figure 5–50 shows an application of a TR switch in connection with a coaxial line. The TR tube is mounted in a cylindrical cavity to form the TR switch. The tube contains two conical electrodes which act as a spark gap. They are brought out through the glass envelope and connected to the cylindrical mount to complete the resonant cavity (lightly shaded area).

A simplified sketch of the resonant cavity with an approximate equivalent circuit is shown in figure 5–51. The coupling loops which are used to couple RF energy into and out of the cavity are represented as transformer windings. The resonant cavity is represented as two resonant circuits in parallel. If the gas at *G* does not break down, the input voltage is transformed to a high level by one tuned circuit and reduced to its original level in the secondary winding at the output. Thus the cavity

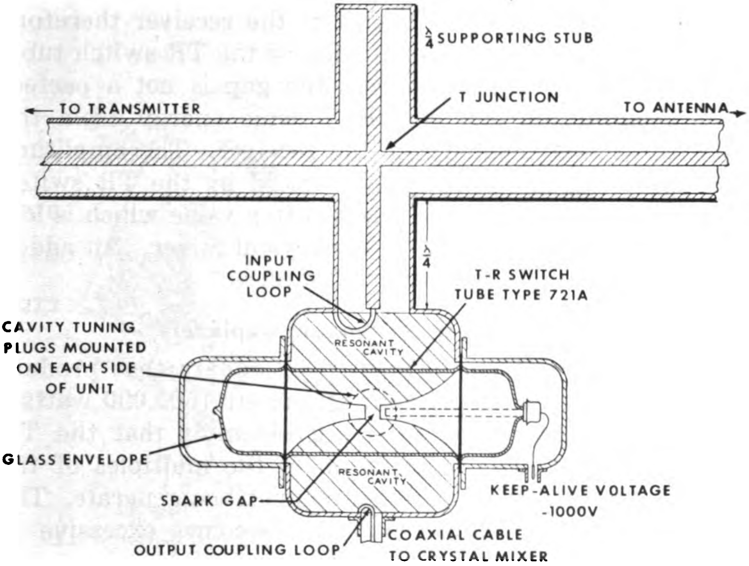


Figure 5–50.—TR switch and coaxial line.

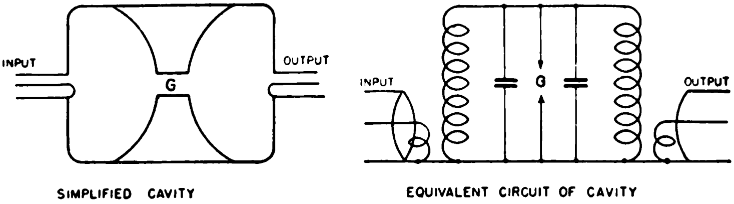


Figure 5–51.—Equivalent circuit diagram of resonant cavity.

acts as a highly selective transformer with a 1:1 turns ratio. If the input voltage is large enough, it is transformed to a voltage sufficient to break down the spark gap. When the gap conducts, the resonant tank is short-circuited, and a short circuit is therefore reflected into the input transformer primary.

Although this conduction does not constitute a perfect short circuit, it causes a very low impedance to appear at the input coupling loop. This very low impedance reflects to the T-junction a quarter wavelength away as a high impedance. The circuit to the receiver therefore is disconnected by the conduction of the TR switch tube.

Since the impedance across the gap is not a perfect short circuit during the pulse, some energy from the transmitter is coupled into the receiver. The amplitude of the voltage so coupled is reduced by the TR switch and transformer stepdown action to a value which is low enough to prevent harm to the crystal mixer. An added safeguard is the KEEP-ALIVE VOLTAGE.

High Power Waveguide Duplexers

A special problem is created by radars that produce outputs in excess of one-half megawatt (500,000 watts). A contributing factor to this problem is that the TR cavity affords little attenuation to the multiples of the carrier frequency which most magnetrons generate. The leakage power of these multiples becomes excessive in high power radars. In order to protect the crystal it becomes necessary to use a modified TR system.

The method generally used to provide the necessary added protection is that of inserting a pre-TR tube at the T-junction ahead of the TR tube. This is illustrated in figure 5–52.

When the TR tube fires, a high impedance is reflected to the input side of the pre-TR which is three-quarter wavelength from the TR tube. This high impedance causes a high voltage drop in the waveguide at this point which causes ionization across the input. This ionization permits the transmitter power to flow with negligible loss and effectively disconnects the receiver from the waveguide. During periods of reception this modified duplexer functions similarly to those previously discussed in this chapter.

Hybrid Duplexer

A type of duplexer somewhat different from those already discussed is the HYBRID RING. This duplexer is capable of performing the duties of the conventional ones already discussed, and in addition it has a low standing wave ratio and the ability to transfer high power.

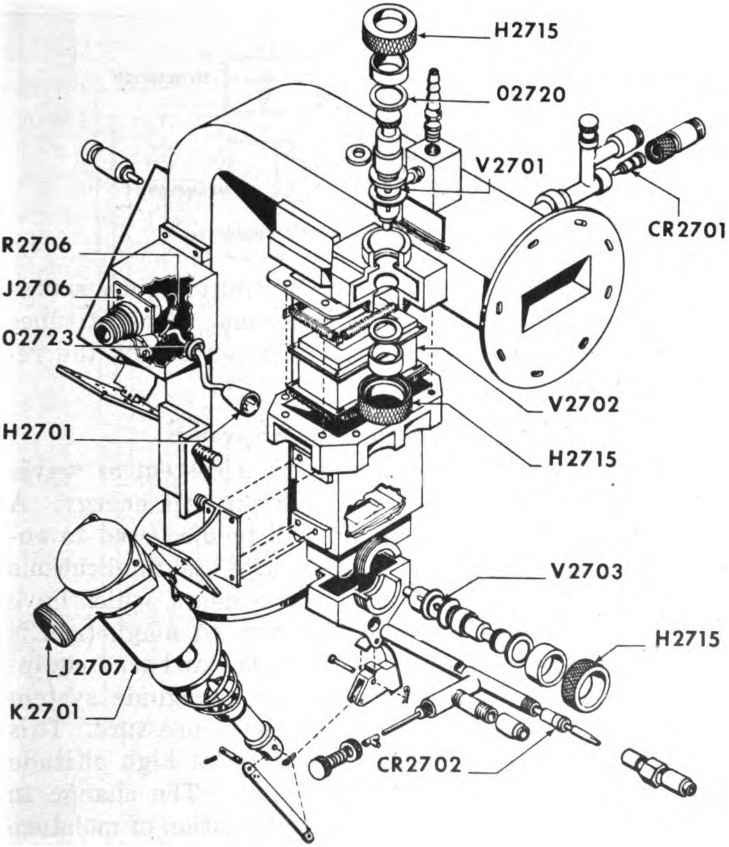


Figure 5–52.—A high power duplexer system.

The hybrid ring shown in figure 5–53 is a form of hybrid junction, which is a device that transmits power to two lines from each of two independent inputs that are not coupled together. This type of junction is frequently referred to as a "magic T." NOTE: This statement holds true for the hybrid ring duplexer only during reception; thus the magnetron is not coupled to the antenna and there is no dissipation of received power by the magnetron.

During the transmitter pulse the duplexer does not function as a true hybrid ring because the ionization of

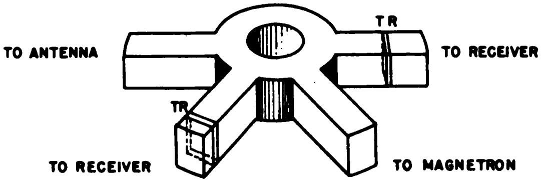


Figure 5–53.—A hybrid ring duplexer.

the TR tubes by the high power transmitter pulse serves to couple the magnetron to the antenna. The TR tubes also perform their normal protective action for the receiver crystal.

A Complete Waveguide RF System

Airborne radar equipments utilize a system of waveguides for the purpose of transmitting RF energy. A typical radar waveguide system will be discussed to enable you to understand the relationships and applications of the waveguide and its special components, which have already been presented in this chapter. A magnetron is generally used as a signal source. Since airborne equipment is flown at high altitudes, the waveguide system must be maintained at least at sea level pressure. This is to prevent arc-over which may occur at high altitude due to decreased atmospheric pressure. The change in atmospheric pressure may cause condensation of moisture in the waveguide and this in turn may cause RF loss, corrosion, and increase the possibility of arc-over. Most aircraft radars use an automatic pressurization system. This system supplies dry air to the waveguide and in most cases to the transmitter and modulator. The air is dried by passing it through a dehydrator before forcing it into the system.

Figure 5–54 shows a waveguide system of an airborne radar installation. The source of RF energy is a magnetron which is connected to the waveguide by the flexible section of waveguide *Z*1305, shown in the upper right-hand corner of the figure. The mode of propagation in this guide is $TE_{0,1}$. During operation, the transmitter out-

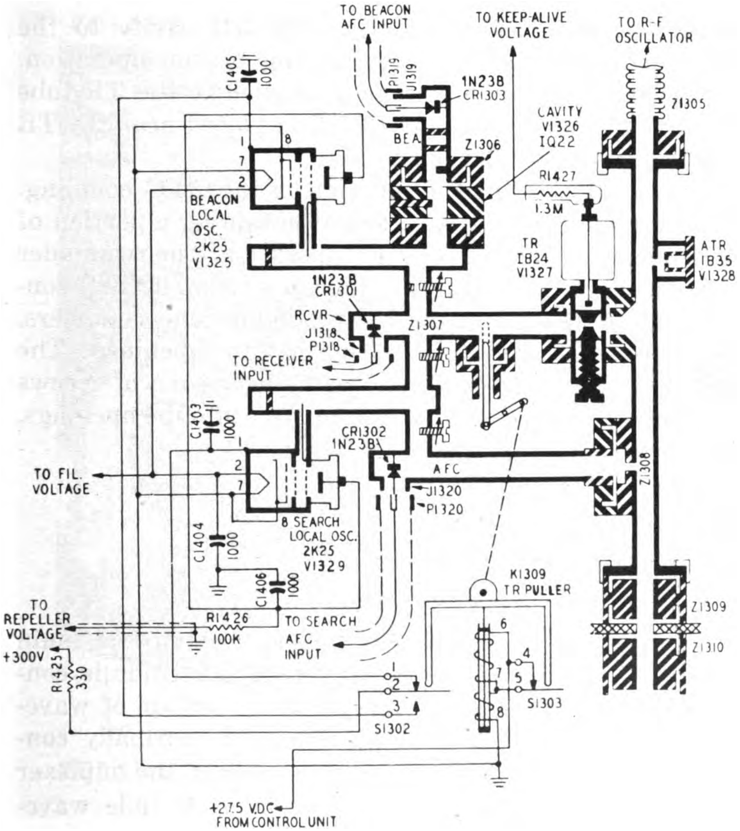


Figure 5–54.—Mixer, duplexer, local oscillators, and associated plumbing.

put in traveling through the waveguide to the antenna passes through the duplexer section. The ATR tube is located at the T-joint nearest the transmitter. The TR tube is located at the second T-connection from the transmitter. In addition to its function as a TR it also provides desirable frequency discrimination and image rejection since it is tuned to the transmitter frequency. Since the TR cavity is tuned to resonate at the transmitter frequency for search radar reception, it presents a bad match for beacon signals. However, during beacon operation the TR puller relay is energized and inserts a plunger into the receiver section of the duplexer. This plunger is adjusted to resonate the TR cavity to the beacon transmitter frequency during beacon operation. Note that a keep-alive voltage is applied to the TR tube and that the series resistor *R*1427 is placed near the TR tube to prevent oscillation.

Located at the third T-junction is the AFC coupling. This junction provides a means of obtaining a portion of the transmitter output for AFC operation. The remainder of the waveguide assembly is known as a mixer and consists of a rectangular block containing four chambers. These are interconnected through narrow openings. The amount of coupling can be adjusted by means of screws (waveguide attenuators) which project into the openings. The four chambers are:

Search local-oscillator.

Beacon local-oscillator.

Receiver-mixer and TR.

Automatic frequency control.

The flow of energy is continuous through the main guide to the antenna. Each section of the guide is connected to the next with a choke joint. A section of waveguide containing a directional coupler is normally connected in series with the waveguide between the duplexer and the antenna assembly. A section of flexible waveguide will most likely be used between the duplexer and the antenna. This serves to insulate the units from mechanical vibrations of the aircraft by providing a flexible connection between shock mounted components.

The antenna shown in figure 5–55 is designed to rotate horizontally and to tilt vertically. Thus, the signal must pass through the two rotating joints before reaching the antenna proper. The radiating system consists of a horn type radiator and a parabolic reflector with a narrow deflecting strip, designed to produce the desired beam (cosecant-squared) pattern. The energy is directed by the horn into the modified parabolic surface from which it is projected forward as a vertical, fan-shaped beam.

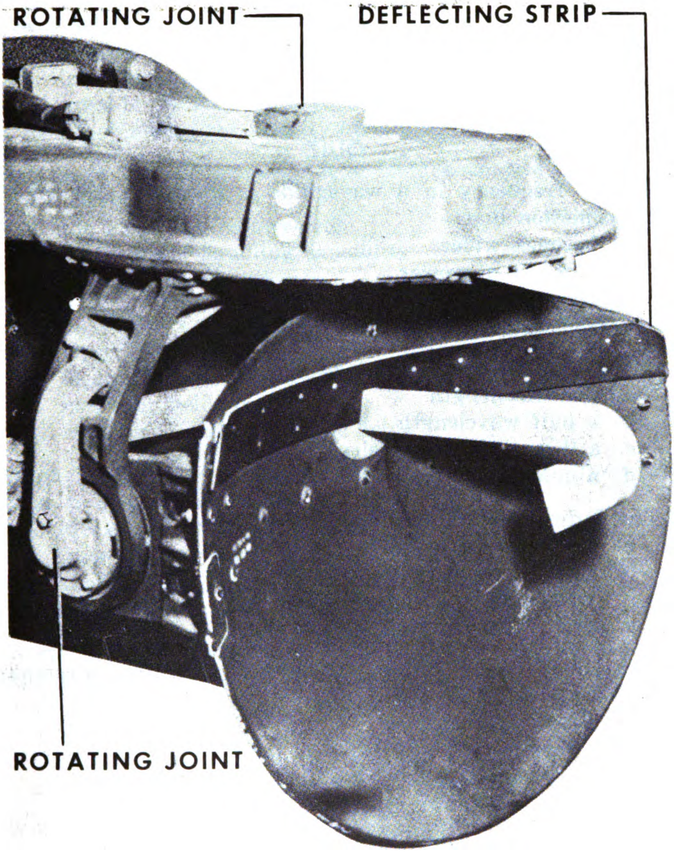


Figure 5–55.—A complete radar antenna assembly.

The energy that returns from the target is gathered by the reflector and concentrated at the horn. It passes from the horn into the waveguide, travels toward the magnetron, and is stopped at the ATR box. It is then reflected back along the guide to the mixing chamber.

QUIZ

1. As the frequency in a waveguide is increased, the effective transmission line
   - a. becomes wider
   - b. becomes narrower
   - c. does not change
   - d. shorts the energy to ground

2. The *B* dimension, at frequency of cutoff, equals
   - a. one wavelength
   - b. a half wavelength
   - c. a three-quarter wavelength
   - d. a quarter wavelength

3. The *A* dimension mainly determines the
   - a. frequency of cutoff
   - b. mode of operation
   - c. power-handling capability
   - d. size of the *B* dimension

4. A disadvantage of the two-wire transmission line, as compared with a waveguide, is the
   - a. larger size required
   - b. higher radiation losses
   - c. lower radiation from the line
   - d. lower breakdown voltage

5. An advantage of waveguide over the coaxial line is
   - a. lower power-handling capability
   - b. lower copper losses
   - c. lower radiation losses
   - d. greater heating of the dielectric

6. To exist in a waveguide, the H-lines must
   - a. be parallel to the E-lines
   - b. be perpendicular to the *B* dimension
   - c. be perpendicular to the *A* dimension
   - d. form complete loops

7. The velocity of propagation of RF energy in a waveguide is
   a. greater than the free space velocity
   b. less than the free space velocity
   c. equal to the free space velocity
   d. greater than the velocity of light

8. A disadvantage of a waveguide at low frequencies is
   a. lower efficiency
   b. the large size requirement
   c. lower power-handling capability
   d. greater radiation losses

9. When a probe is used to excite a waveguide, it sets up
   a. electrostatic fields
   b. electromagnetic fields only
   c. magnetic fields only
   d. electrostatic and magnetic fields

10. What is the wavelength required to make a 90-degree bend so as to avoid reflections?
   a. Greater than 3 wavelengths
   b. Less than 2 wavelengths
   c. Greater than 2 wavelengths
   d. Less than 3 wavelengths

11. What mode of operation must be used in a waveguide acting as a rotating joint?
   a. $TM_{1,1}$
   b. $TE_{1,1}$
   c. $TE_{0,1}$
   d. $TM_{0,1}$

12. When plugs are inserted in the cylindrical side of a cavity, the deeper the plug extends into the cavity, the
   a. higher the frequency
   b. lower the frequency
   c. smaller the magnetic field strength
   d. smaller the electrostatic field strength

13. A keep-alive voltage is applied to the
   a. pre-TR tube
   b. ATR tube
   c. TR and ATR tubes
   d. TR tube

14. One function of the TR tube is to
    a. keep most of the RF energy from entering the radar receiver during transmit time
    b. direct the RF energy into the transmitter section of the radar
    c. prevent an output from the radar transmitter during the receive period
    d. add the necessary impedance to the duplexer system during transmit time

15. The main function of the ATR tube is to
    a. prevent RF energy from leaving the transmitter during transmit time
    b. prevent received RF energy from getting back into the transmitter section
    c. compensate for all losses within the duplexer section
    d. trigger the transmitter at the proper time

16. Recovery time may be described as the time required for a TR tube to
    a. become ionized
    b. become de-ionized
    c. reach its operating temperature
    d. become completely cold

17. In a properly operating duplexer system, during transmit time, the ATR tube
    a. and the TR tube are in an unfired state
    b. is unfired and the TR tube is fired
    c. is fired and the TR tube is fired
    d. is fired and the TR tube is unfired

18. In high-power type duplexer systems, a device used to prevent excessive power leakage to the radar receiver during transmit time is called a
    a. cavity attenuator
    b. pre-TR tube
    c. ATR tube
    d. silicon crystal

CHAPTER 6

UHF AND MICROWAVE TECHNIQUES

The purpose of this chapter is to acquaint you with oscillators and oscillatory circuits employed at UHF and microwave frequencies. Also, information dealing with the pulsing techniques used in RF generators of this frequency will be presented. Specifically, it discusses problems involved in generating high-frequency signals, explains the theory and operation of a number of oscillators employed at these frequencies, and describes various methods for pulsing oscillators.

ADVANTAGES OF HIGH FREQUENCIES

Perhaps the main reason for using higher frequencies is that it is possible to construct directive antennas of smaller size than when lower frequencies are used. Small directive antennas are an important consideration in airborne equipment where space and weight are major problems.

Aside from the factors of size and weight, these frequencies are highly applicable to radar equipment for they produce better target echoes and make possible the detection of smaller targets. The size of the target that radar can detect depends mostly on the frequency. In general the higher the frequency, the smaller the target that the radar can detect. A target gives the best echoes when it is about a half wave or some multiple of the radar frequency being transmitted. If the target is much smaller than a half wave, the intensity of pulse that it reflects is low. However, as the operating frequency is increased, the half wave becomes smaller and makes it possible to detect smaller targets.

An overall statement as to an advantage of high frequencies is that the trend toward miniaturization becomes a reality. This is highly desirable for aviation equipment. Circuit components generally become smaller as the frequency increases.

FREQUENCY LIMITATIONS OF CONVENTIONAL OSCILLATORS

The principal factors which limit the frequency of the ordinary type oscillator are the construction of the tube and the external circuits which connect to the tube.

Tube Construction and Limitations

The tube construction factors which limit the operating frequency of the conventional oscillator are the interelectrode capacitances in the tube, the inductances of the leads, and the transit time.

INTERELECTRODE CAPACITANCES.—At low or medium radiofrequencies the interelectrode capacitances in an ordinary vacuum tube have reactances which are so large that they do not cause any serious trouble. However, as frequencies increase, the reactance of these capacitances become small enough to materially affect the performance of a circuit. A 1-μf. capacitor, for example, has a reactance of 1,590 ohms at 100 mc. If this capacitor is the grid-to-plate capacitance and the RF voltage between these electrodes is 500 volts, there will be an interelectrode capacitance current flow of 500/1,590 or 0.315 amperes, an amount of current which will disturb circuit operation. On the other hand, at 1 mc. the reactance of this capacitor becomes approximately 159,000 ohms and the current flow is only 500/159,000 or 3.15 milliamperes, an amount which will not seriously affect circuit performance.

A good point to remember is that the higher the frequency or the larger the interelectrode capacitance, the higher the current flow through this capacitance. In most

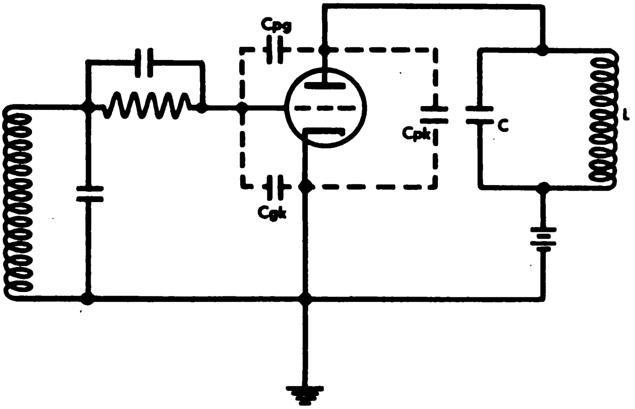


Figure 6–1.—Effect of interelectrode capacitance on frequency of tuned-plate tuned-grid oscillator.

UHF oscillators employing ordinary tubes, interelectrode capacitance currents are much greater than the power currents supplied by the tube. These higher currents cause power losses in the resistance in the oscillatory circuit.

Since interelectrode capacitances are effectively in parallel with the tuned circuit, they affect the frequency at which the tuned circuit resonates. As you can see in the circuit illustrated in figure 6–1, the plate-to-cathode capacitance is in parallel with the series combination of the plate-to-grid capacitance and the grid-to-cathode capacitance. All these capacitances together form a part of the total capacitance of the plate's tuned circuit. You can find their total capacitance by the formula,

$$C_t=C+C_{pk}+\frac{C_{gk}C_{gp}}{C_{gk}+C_{gp}}$$

Not only does interelectrode capacitance limit the frequency by establishing a minimum value below which it is impossible to go, but it also varies with the applied voltages and with the loading of the oscillator. This causes frequency instability, particularly when the interelectrode capacitance forms a large part of the tuning capacitance.

INDUCTANCE OF LEADS.—Another frequency limiting factor within a tube is the inductance of the leads to each tube element. While these inductances do not necessarily impair the efficiency of the oscillator, they may represent a major portion of the inductance of the tuned circuit and limit the frequency by setting a minimum limit on the inductance. Furthermore, since the cathode lead is frequently common to both plate and grid circuits, feedback takes place through it and produces an additional loss of efficiency.

TRANSIT TIME.—A third limitation imposed by tube construction is TRANSIT TIME. Transit time is the time required for electrons to travel from cathode to plate. At low radiofrequencies transit time is negligible, since it occupies only a comparatively small portion of an oscillatory period. But as the frequency becomes higher, transit time occupies an appreciable portion of this period and produces undesirable effects in tube operation. The effect of transit time is of special concern in connection with the input impedance of the tube. Part of the current that flows in the grid circuit is the current which charges the grid-to-plate capacitance, C_{gp}. The voltage that produces this current is the vector sum of the input voltage (grid to cathode) and the output voltage across the plate load.

At lower frequencies with a resistive load, these two voltages are 180 degrees out of phase and add algebraically to determine the charging current. This current is 90 degrees out of phase with the input voltage. However, at higher frequencies, where transit time is an important factor, the plate current begins to lag the input voltage. This causes the plate voltage to be less than 180 degrees out of phase with the input voltage, and the voltage across the capacitor to lag the input voltage slightly. As a result, the charging current will no longer be 90 degrees out of phase but will have an in-phase component. This means that power is consumed in the grid circuit. This consumption of power is effectively the same as adding a high resistance between the grid and cathode. This resistance decreases as the frequency increases.

OVERCOMING TUBE LIMITATIONS.—There are several ways to reduce the effect of interelectrode capacitances in vacuum tubes. None are completely satisfactory. One is to move the electrodes farther apart. However, this is hardly desirable for it increases the transit time. Another method is to reduce the size of the tube and electrodes. This is satisfactory except for one ill effect. It decreases the power handling ability of the tube directly with the square of the factor by which the electrodes are reduced. Another method is to separate the leads and to bring them out of the envelope at the nearest point. This results in a slight decrease in the capacitances.

Similarly there are several ways to reduce the inductances of the leads. As just mentioned, bringing out electrode leads through the envelope at the nearest point produces a slight decrease in the electrode capacitance in a tube. This also decreases the inductances of the leads. Another method is to make double connections to the electrode. This makes two parallel inductances which cut the lead inductance in half. Another method is to arrange the leads as extensions of the external transmission lines which are the resonant elements of the circuit. Thus the interelectrode capacitances and lead inductances are incorporated as part of the tuned circuit. (See fig. 6–9.)

It is possible to minimize the inefficiency previously mentioned by using a tube that has two separate external cathode connections. When one cathode lead is used for the grid return and the other for the plate circuit the unwanted feedback is prevented, since the grid and plate currents do not flow in a common lead inductance. (The 6AG5 and 6AK5, as well as many other tubes, are constructed in this manner.)

For reducing transit time, closer spacing of the electrodes and higher plate voltage are employed. Closer spacing of electrodes, however, causes higher interelectrode capacitances. Therefore, it can be used only when the electrodes are made smaller. For this reason UHF oscillators operate at very high plate voltage. The fact that a radar and other pulsed transmitters do not operate continuously is an advantange in that high enough voltages may be used to reduce transit time without exceeding the maximum power rating of the tube. Operating with high voltage does, however, necessitate some precautionary measures, such as separating the leads to avoid the presence of excessive voltage gradients in or on the surface of the glass envelope, and avoiding sharp projections on the high-voltage leads or electrodes to prevent arcing.

Limitations Due to External Circuits

Increases in frequency cause a rather rapid increase in the power losses in the external oscillatory circuit. These power losses are due principally to skin effect and radiation.

SKIN EFFECT.—Skin effect is the characteristic of conductors at radiofrequencies in which most of the current flows near the surface of the conductor. The depth below the surface where the density of current flow is reduced to 36.8 percent of the surface value (in a copper conductor) is equal to 6.62 divided by the square root of the frequency in cycles. (This will be the depth in centimeters.) For example, this depth would be 0.0066 cm. for one megacycle and 0.00066 cm. for 100 megacycles, thus the depth at 100 megacycles is one-tenth as great as at one megacycle.

Skin effect causes a considerable increase in the resistance in a vacuum tube circuit. This results in a lower Q and increased I^2R losses. To prevent skin effect losses, conductors are made large in size and tubular in shape since current flows only in the surface at UHF frequencies. In addition they are plated with silver, since it has a higher conductivity than copper. At UHF frequencies most and possibly all current flow will be in the silver plating.

RADIATION.—Radiation, the other cause of power loss, is due to incomplete cancellation of electromagnetic fields in the region surrounding the circuit. When the frequency is low enough so that the spacing between two parallel conductors equals only a very small fraction of a half wavelength, there is almost complete cancellation of fields in all directions. At higher frequencies, however, the same spacing would represent a larger fraction of a half wavelength. This means less cancellation. In extreme cases, such as where the spacing is a half wavelength, the fields add in the direction of the plane containing the two conductors. This causes the tuned circuit to radiate energy like an antenna. Therefore, as the frequency increases it is necessary to reduce the spacing between the parallel elements. However, there is a limit on how far you can go in reducing the spacing; too close spacing will result in arcing of the tank circuit. Another means of eliminating radiation is the use of concentric lines instead of open wire line. This eliminates radiation entirely, since the outer conductor acts like a shield which prevents the electromagnetic field from expanding into space.

LIMIT OF INDUCTANCE AND CAPACITANCE.—At UHF frequencies it is necessary that the inductances and capacitances in the oscillatory circuit be very small. The lower limit of capacitance is the sum of the interelectrode capacitances and the distributed capacitance of the leads. The lower limit of inductance is the lead inductance plus the inductance necessary to connect the tube electrodes externally. UHF oscillators approach both these limits, the only capacitance in the tank circuit often being a small trimmer capacitor for fine tuning adjustments, and the inductance often being a short-circuited transmission line less than one-quarter wavelength in length. In the typical parallel line tank circuit (fig. 6–2), the expression for the input (sending end)

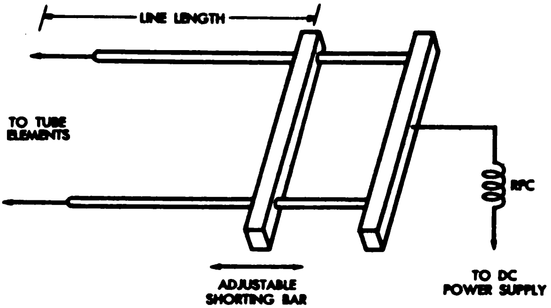


Figure 6–2.—Typical parallel wire resonant circuit.

reactance X_s (looking at the circuit from the tube) of the tank circuit is

$$X_s = {}_jZ_o \tan B_s$$

where Z_o is the characteristic impedance of the line and B_s the electrical length.

As you can see in the expression, if B_s is less than 90 degrees ($\lambda/4$), then tan B_s is positive and the reactance is inductive. For smaller values of inductance the length of the line is decreased. Further, this expression is true only when the resistance losses are low enough to be neglected. Frequently the tank circuit is located near a flat surface and parallel to it. The flat surface acts as a shield and, while it does not affect the electrical length of the line, it does lower the characteristic resistance of the line.

TRIODE OSCILLATORS

Ultraudion Oscillator

The typical ultraudion oscillator illustrated in figure 6–3 is sometimes used as a local oscillator in receivers and occasionally as a transmitter. The d-c circuits contain the components R_p, C_g, R_g, and the small d-c resistances in the tank circuit and leads. The components C_g and R_g form the grid leak bias circuit. The charge built up on C_g biases the tube to class C operation.

To understand the operation of the ultraudion oscil-

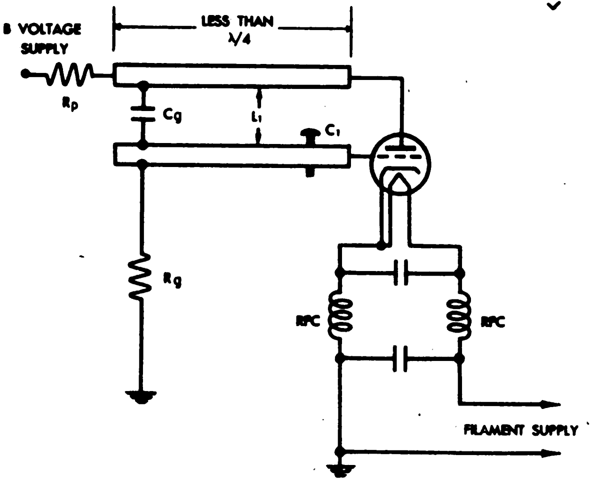


Figure 6–3.—Ultraudion oscillator.

lator, consider its equivalent circuits shown in figure 6–4. So far as RF is concerned, as you can see in the equivalent circuit, the chokes in the cathode are open circuits. The tank circuit inductance L consists of the plate and grid lead inductances connected in series with the parallel combination of L_1 and C_1.

The simplified equivalent circuit shows that this tank circuit inductance forms a resonant circuit with the combined interelectrode capacitance, C, which is given by the equation

$$C=C_{gp}+\frac{C_{gk}\times C_{pk}}{C_{gk}+C_{pk}}$$

The expression for the frequency of this resonant circuit is

$$f_o=\frac{1}{2\pi\sqrt{LC}}$$

Basically the circuit of the ultraudion oscillator, as you can see in the simplified equivalent is that of a Colpitts oscillator. The capacitances C_{gk} and C_{pk} form a voltage divider circuit. The voltage E_g developed across

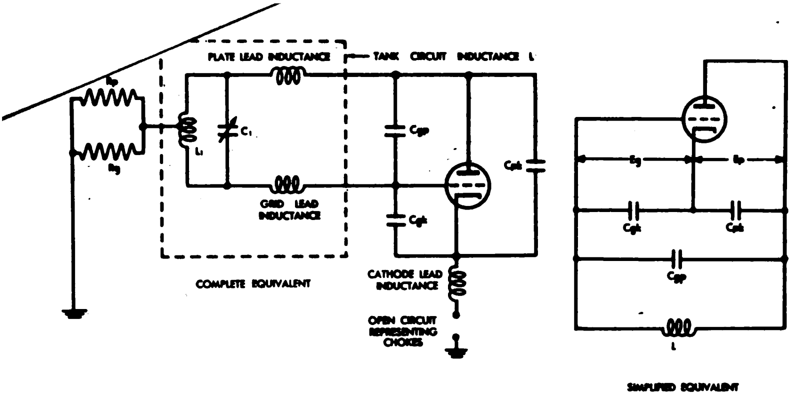


Figure 6–4.—Equivalent circuits of ultraudion oscillator for RF currents.

C_{gk} is in the right phase and high enough to sustain oscillations. The minimum amplification required to sustain oscillations is given by the formula

$$A=\frac{C_{gk}}{C_{pk}}$$

If the initial gain is greater than this value, oscillations build up to such an amplitude that the gain is reduced to the value given by this formula.

Lighthouse-Tube Oscillator

The lighthouse-tube oscillator like the ultraudio oscillator is a single tube oscillator. It is used in test equipment and low-power transmitters. The lighthouse tube, which gets its name from its peculiar construction, will operate at frequencies as high as 4,000 mc. Its plate, grid, and cathode connections are arranged so that it may be mounted directly in the tuning assembly. (See fig. 6–5.) The component called the tuner consists of three coaxial cylindrical conductors, of which the inner cylinder makes contact with the plate, the next with the grid, and the outer conductor with the shell of the tube. The outer conductor provides an RF connection to the cathode through capacitance. The space between the cathode and grid conductors forms a coaxial-cathode line which in turn is shorted by means of a plunger. The

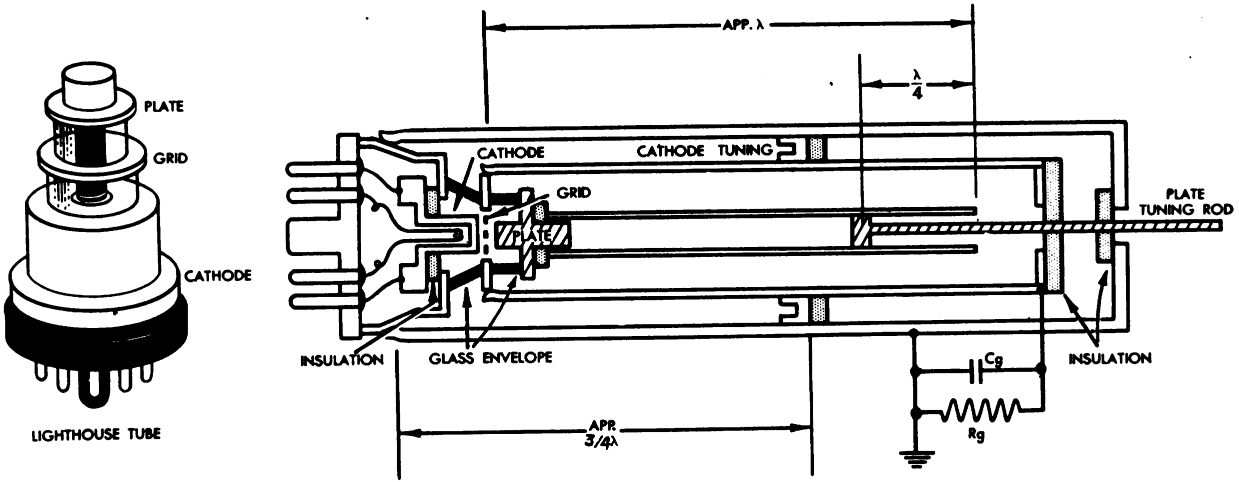


Figure 6–5.—Lighthouse-tube oscillator.

insulation at the tube base between the cathode and the cathode shell (or external connection) is the dielectric between the rings which form the cathode capacitive coupling. If this capacitive coupling were not used, the grid would have a d-c path from its cylinder through the plunger and the cathode cylinder to the cathode. This capacitance in the tube base is great enough to form a virtual short circuit for RF. The d-c grid-to-cathode path consisting of R_g and C_g form the grid leak bias circuit.

In the plate circuit, the plate line consists of the grid and plate conductors. It is open circuited at the end away from the tube. The plate voltage is applied to the plate lead through the plate tuning rod at the point where the plate line is shorted. This point is one-quarter wavelength from the end. This quarter-wave section which is shorted at the point where B plus is applied presents a high impedance to RF from the open-ended plate line and therefore prevents RF from flowing into the power supply.

EQUIVALENT CIRCUITS.—The equivalent circuits in figure 6–6 show the evolution of the lighthouse-tube oscillator. In the first equivalent, (A), the components L_k and L_p represent the inductances of the shorted cathode line and the open circuited plate line respectively. This is a true representation of the actual circuit provided that the cathode line is less than three-quarter wavelength long and provided that the plate line is less than a full wavelength long. The three-quarter wavelength and full wavelength were chosen for this circuit rather than the one-quarter and one-half wavelength due to the fact that at the frequency of operation the latter fractions of a wavelength are inconveniently short. However, keep in mind that if the cathode line is slightly less than one-quarter wavelength, it acts like an inductance and that if the plate line is slightly less than one-half wavelength, it will act like an inductance. There is no lead inductance shown in the circuit since the lines

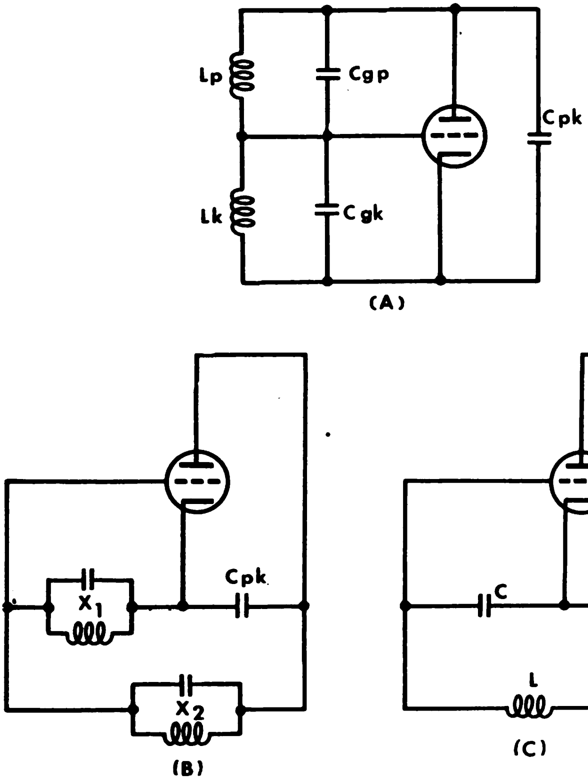


Figure 6–6.—Evolution of equivalent circuit of lighthouse-tube oscillator.

connect directly to the electrodes. The interelectrode capacitances complete the equivalent circuit.

In the equivalent circuit (B), the parallel combination of L_p and C_{gp} are represented by X_2 and the parallel combination of L_k and C_{gk} by X_1. Obviously the reactances X_2 and X_1 along with C_{pk}, which you can consider unknown reactances, must form a resonant circuit. The reactance X_1 forms a voltage divider circuit along with C_{pk}. Since the voltage across X_1 is the voltage fed back to the grid, it must be 180 degrees out of phase with the plate voltage which can happen only if X_1 is a capacitive reactance. This condition leads to the third equivalent circuit (C), in which the reactance X_1 is represented by a capacitor.

An oscillatory circuit requires at least one inductance, therefore X_2 must be an inductive reactance. Hence, in the final or the third equivalent circuit (C), the reactance X_2 is shown as the inductance L. This circuit is representative of the oscillator circuit only at the resonant frequency. On examining this circuit, notice that basically it is that of a Colpitts oscillator. Now, if X_1 is capacitive, C_{gk} must conduct more heavily than L_k; therefore the oscillator frequency must be above the resonant frequency of L_k and C_{gk}. Likewise, for X_2 to be inductive the oscillator frequency must be below the resonant frequency of L_p and C_{gp}. Thus you see that the oscillator frequency is between the resonant frequencies of the plate circuit and the cathode circuit.

FEEDBACK.—In this oscillator, the amount of feedback depends upon the size of C in relation to C_{pk}. This is, in turn, a function of the tuning of the cathode line; hence, the cathode tuning serves principally for controlling the amount of feedback. If feedback is too small, oscillations will be weak and may cease completely. If feedback is too great, power consumption in the grid circuit will become too large and will result in a decrease in output of the oscillator. The value of C, hence of the cathode tuning, has some effect on the frequency of oscillations, but since C and C_{pk} are in series and C is usually much larger than C_{pk}, C must be changed considerably to affect the frequency appreciably.

PLATE TUNING.—Since the tuning of the plate line determines the resonant frequency of the plate circuit, which in turn determines the amount of inductance represented by the plate circuit at the oscillator frequency, the plate tuning is the chief control of the oscillator frequency. It has a secondary effect of feedback since C is determined by the relative frequencies of the cathode circuit and of the oscillator. You can see from the foregoing discussion that the plate line is tuned to give the correct frequency and the cathode line to adjust feedback for maximum output from the oscillator but, since there is some interaction, it will be necessary to readjust each line for maximum accuracy.

Push-Pull Oscillators

To obtain higher power outputs, two tubes in push-pull circuits or four or more tubes in a ring oscillator are sometimes used. In push-pull circuits, the tuned circuits are the parallel line (Lecher line) type. This arrangement forms a balanced-to-ground system.

The oscillator circuit discussed first is the push-pull tuned-plate tuned-grid oscillator. This circuit may oscillate continuously or it may be used to provide pulsed operation; this will be explained later in this chapter.

TUNED-PLATE TUNED-GRID OSCILLATOR.—In the tuned-plate tuned-grid oscillator, both the plate and grid lines are slightly under $\lambda/4$ in length. They are indicated in the circuit of the tuned-plate tuned-grid oscillator (fig. 6–7) as inductances L_p and L_g. The length of the plate line in the circuit may be varied by the shorting bar. While the grid line is of fixed length in this particular arrangement, a little greater flexibility would result if it too were adjustable in length. The inductances L and capacitors C_1, C_2, C_3, and C_4 in the cathode circuits form filters to prevent RF from reaching the filament transformer. The grid leak bias developed includes R_g and C_g. The component C_g is shown as a dotted portion since it is actually the distributed capacity of the grid line to ground. B-plus voltage is applied to the plate lines through R_s and L_s. A double purpose is served by L_s: First, it prevents RF from getting into the power supply; and second, it suppresses parasitic oscillations.

EQUIVALENT CIRCUITS.—In equivalent circuit (A), figure 6–8, both tubes of the oscillator are shown. L_p is the inductance, which represents both the plate line and the plate lead inductances. L_g represents the inductance of the grid line and grid leads. L_k is the cathode lead induct-

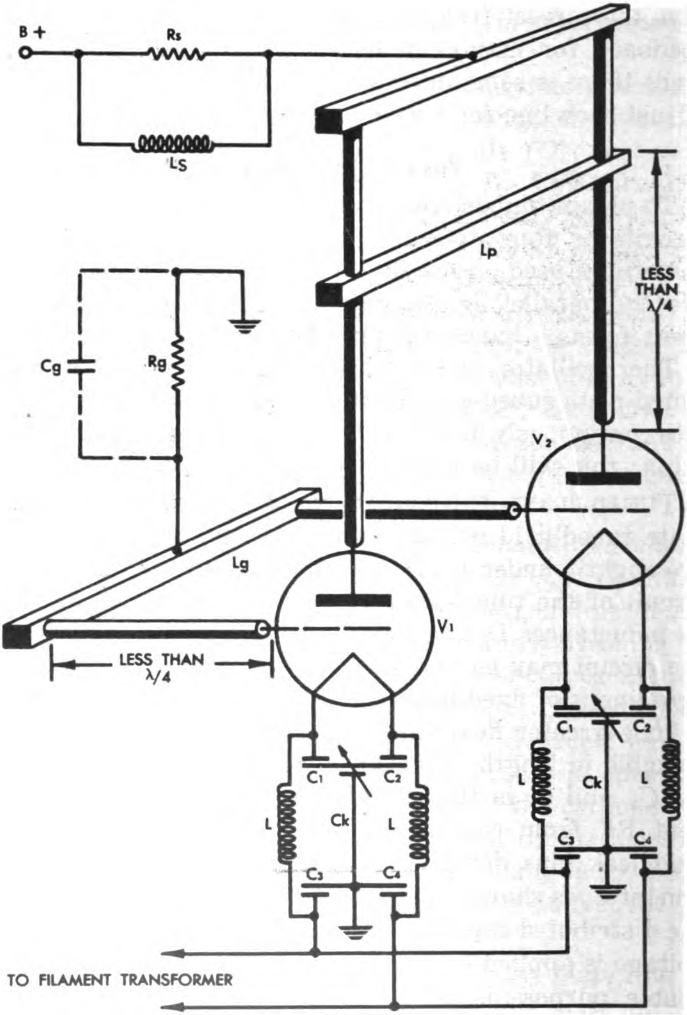


Figure 6–7.—Tuned-plate tuned-grid oscillator using Lecher lines as tuned circuits.

ance. With proper adjustment of C_k, L_k forms a series resonant circuit that effectively grounds the cathode within the tube.

The second equivalent circuit (B) shows only one tube since both circuits are identical. L_{pk} is one-half L_p and

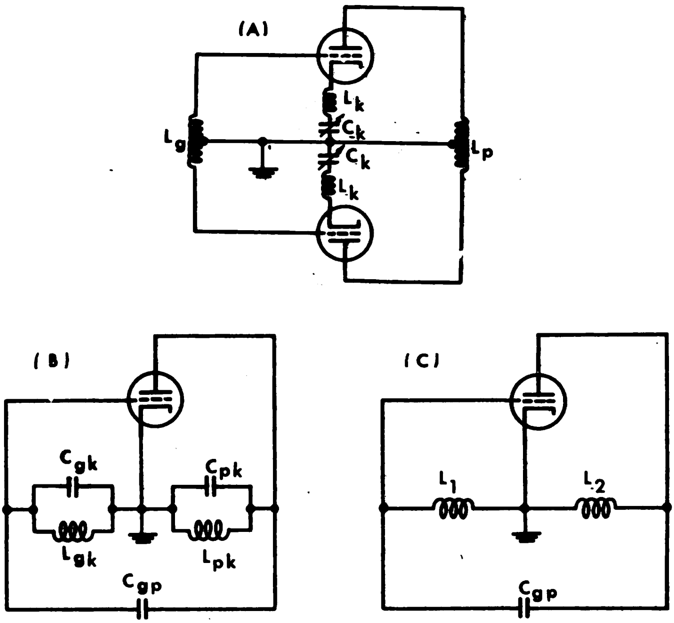


Figure 6–8.—Equivalent circuits of TPTG oscillator.

along with C_{pk} forms a circuit that resonates at a frequency above that at which the entire circuit oscillates.

In the third equivalent circuit (C), the parallel combination of L_{pk} and C_{pk} is represented as an inductance L_2. Likewise L_{gk}, which is one-half L_g and along with C_{gk} forms a circuit which resonates at a higher frequency than that of the oscillator, is represented as an inductance L_1 at the oscillator frequency. This is true because the tank circuit must have at least one inductive element to oscillate and the two reactances must be alike for the feedback to be in the correct phase. In this equivalent circuit, the oscillator is the Hartley type. Since the oscillator frequency must be lower than either resonant frequency (plate or grid circuits), the grid or plate circuit, whichever is lower in resonant frequency, controls the oscillator frequency and the other controls the amount of feedback.

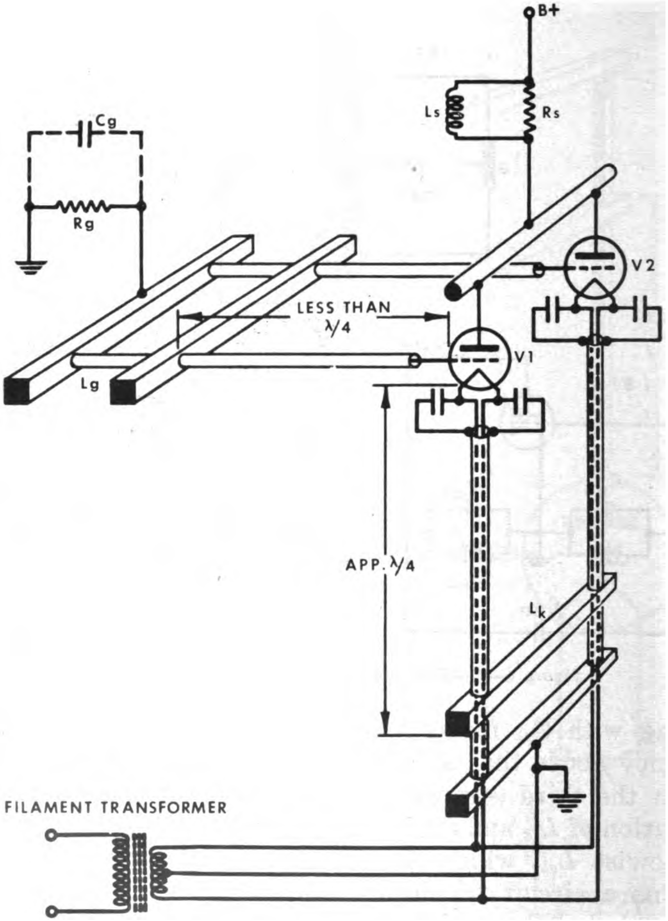


Figure 6–9.—Tuned-grid tuned-cathode oscillator.

TUNED-GRID TUNED-CATHODE OSCILLATOR. — Another push-pull oscillator circuit which is used in a number of types of UHF equipments is the tuned-grid tuned-cathode circuit (fig. 6–9). In this oscillator, the plates are effectively short-circuited by tubes which have very little plate lead inductance and which are connected by a section of line. When the lead inductance is large enough to become objectional, the plates are connected by means of a half-wave section of line to insure a short circuit.

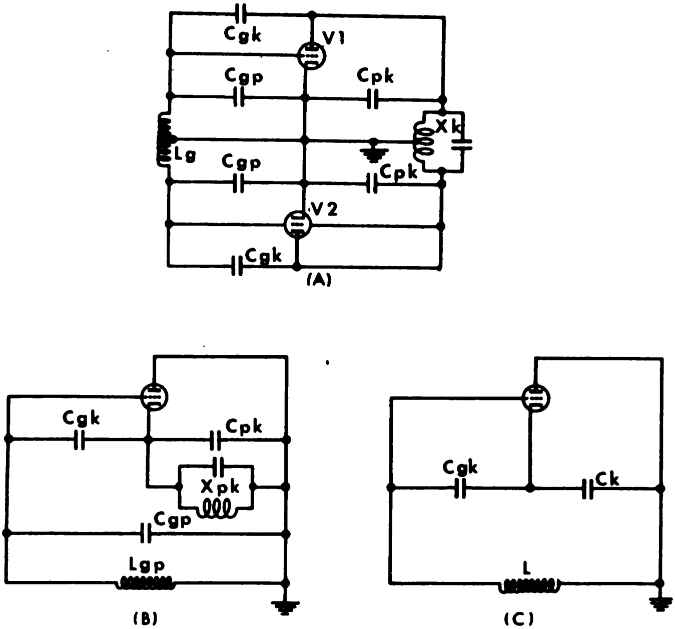


Figure 6–10.—Equivalent circuits of tuned-grid tuned-cathode oscillator.

The d-c circuits are the same as in the tuned-plate tuned-grid oscillator.

EQUIVALENT CIRCUITS.—In the first equivalent circuit (A), figure 6–10, both tubes are shown. The grid line is represented as inductance L_g, the cathode line as reactance X_k, and the interelectrode capacitances as indicated.

Circuit (B) shows only one tube. In this circuit L_{gp} represents one-half of L_g and X_{pk} represents one-half of X_k. By following the line of reasoning similar to that used in the TPTG circuit, you see that the parallel combination of X_{pk} and C_{pk} must be capacitive at the oscillator frequency. Thus it is shown as capacitor C_k in the third equivalent circuit (C).

The oscillator must have an inductive element; hence L_{gp} and C_{gp} in parallel act like an inductance L as shown in (C). For L_{gp} and C_{gp} to be inductive, the grid line must always be less than $\lambda/4$ in length. Since variations in L

have the greatest effect on the frequency of oscillations, THE GRID LINE IS THE PRIMARY FREQUENCY TUNING DEVICE. The cathode line (X_{pk}) controls only a portion of the total capacitance of the circuit and therefore does not greatly affect the oscillator frequency. However, since X_{pk} in parallel with C_{pk} forms a voltage divider with C_{gk}, it has a considerable effect on feedback. X_{pk} may be inductive, purely resistive, or capacitive, depending on whether the cathode-line length is less than, equal to, or greater than a quarter wavelength at the frequency of oscillation. If X_{pk} is a high resistance, the feedback is determined entirely by the voltage divider composed of C_{gk} and C_{pk}. If X_{pk} is capacitive (greater than $\lambda/4$), then the two capacitances in parallel combine to increase C_k and results in an increase in feedback. Tuning the cathode line to slightly less than $\lambda/4$ makes it inductive and decreases C_k, which results in a reduction of feedback. Oscillations will cease if X_{pk} and C_{pk} in parallel can no longer be represented as a capacitor C_k. As can be seen by the third equivalent circuit, this oscillator is basically a Colpitts oscillator.

ADVANTAGES.—The tuned-grid tuned-cathode type oscillator has three advantages: First, the cathode line offers a convenient means of connecting the load to the oscillator because it is at ground potential with reference to d.c.; second, since only d-c voltages appear at the plate and only RF voltages appear in the cathode circuit, the two are not present in the same circuit and the high peak voltages present in a tuned-plate circuit are avoided; and third, the tuning of the oscillator is quite simple, the grid line being the frequency control and the cathode line the feedback control. There is some interaction between the tuning controls so that readjustment of each may improve both accuracy and output.

Tuned-Plate, Tuned-Grid, Tuned-Cathode Oscillators

A number of oscillators in the UHF range employ tuned lines in all three leads of the triode, as in the oscillator circuit illustrated in figure 6–11. So far as RF is concerned, all three tuned lines are grounded at their

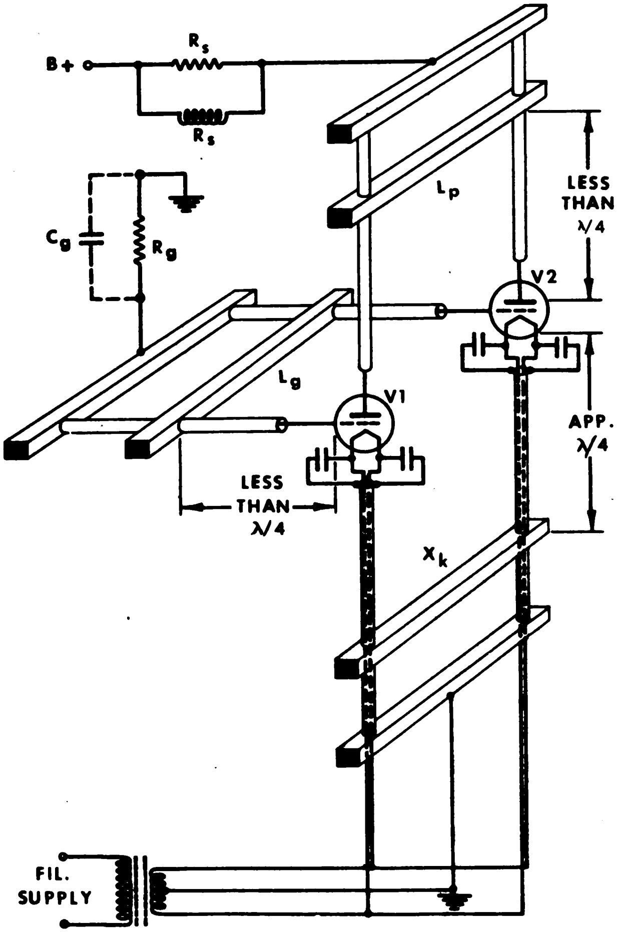


Figure 6–11.—Tuned-plate, tuned-grid, tuned-cathode oscillator circuit.

midpoints. The d-c circuits and filament circuits are identical to those described previously. In the circuit shown, the plate and grid lines, which are less than a quarter wavelength long, are labeled L_p and L_g respectively. The cathode line, which may be equal to, less than, or greater than a quarter wavelength, is labeled X_k.

The equivalent circuit, figure 6–12 (A), shows both tubes, the inductances L_p and L_g center tapped to ground,

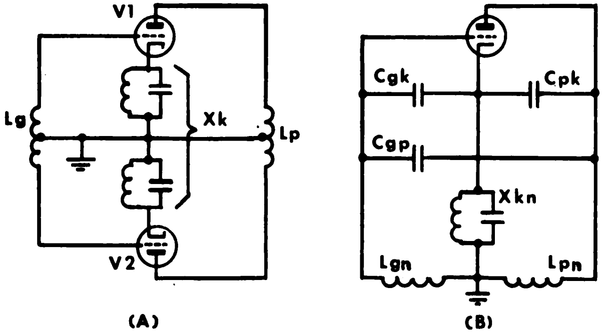


Figure 6–12.—Equivalent circuits of tuned-plate, tuned-grid, tuned-cathode oscillator.

and the unknown reactance X_k center tapped to ground. The interelectrode capacitances are not shown. Since the circuits of the two tubes are identical, the second equivalent circuit (B) shows only one tube. The inductances L_{pn} and L_{gn} and the reactance X_{kn} represent one-half of L_p, L_g, and X_k. The interelectrode capacitances are added. Examination of circuit (B) shows that if the cathode line is exactly a quarter wavelength so that X_{kn} becomes a high resistance, then the circuit is that of a conventional Colpitts oscillator. If the ratio of L_{gn} to L_{pn} is the same as the ratio of the reactances of C_{gk} and C_{pk}, there is no difference in potential between the cathode and ground and X_{kn} is in effect out of the circuit. This condition also reduces the circuit to that of a Colpitts oscillator. In either case, the frequency of oscillation is determined by the sum of the inductances of the plate and grid lines, and the amount of feedback by the ratio of C_{gk} and C_{pk}. Ordinarily this ratio does not give the amount of feedback necessary to cause oscillations; for best efficiency, therefore, the cathode line is detuned from the pure resistive quarter wavelength so that X_{kn} is capacitive or inductive.

Making X_{kn} capacitive or inductive while keeping the sum of L_{gn} and L_{pn} constant does not change the frequency of oscillation but does increase feedback. The exact amount of feedback depends on the ratio between L_{gn} and L_{pn}. Thus, for example, consider the case where L_{pn} is reduced to zero and examine the nature of X_{kn}. In this case X_{kn} is in parallel with C_{pk} as shown in equivalent circuit of the tuned-grid tuned-cathode oscillator in figure 6–10 (B). Thus, in controlling feedback, the effect of X_{kn} changes is the same as in the tuned-grid tuned-cathode oscillator—that is, the cathode line must be increased in length to increase feedback or be decreased in length to decrease feedback.

On the other hand, if L_{gn} is decreased to zero and L_{pn} is increased to equal the value of their former sum, then X_{kn} is in parallel with C_{gk}. Its effect on feedback then is just the opposite to that described. For example, if the cathode line is greater in length than a quarter wave, it becomes capacitive and the parallel combination C_{gk} and X_{kn} becomes a larger capacitance. This reduces the feedback. Reducing the length of the cathode line below a quarter wavelength makes it inductive, and the parallel combination becomes a smaller capacitance. This results in an increase in feedback. This demonstrates that FEEDBACK IS A FUNCTION OF BOTH THE RATIO OF L_{gn} AND L_{pn} AND OF THE TUNING OF THE CATHODE LINE. Since the frequency of oscillations is controlled by the sum of L_{pn} and L_{gn} and feedback is controlled by both the ratio of L_{pn} and L_{gn} and the cathode-line length, there are numerous settings of the three variable line lengths for any given frequency of oscillation and for proper feedback. This is readily understood if you consider that you have three variable elements with which to control two variables, and that there is a certain amount of interaction among the three controls.

Pulsing of Triode Oscillators

There are two methods of pulsing UHF oscillators. One method employs separate modulator tubes; the other employs a self-pulsing oscillator.

THE SELF-PULSING OSCILLATOR.—Most oscillators use grid leak bias or a combination of grid leak and fixed bias. An advantage of grid leak bias is that the ampli-

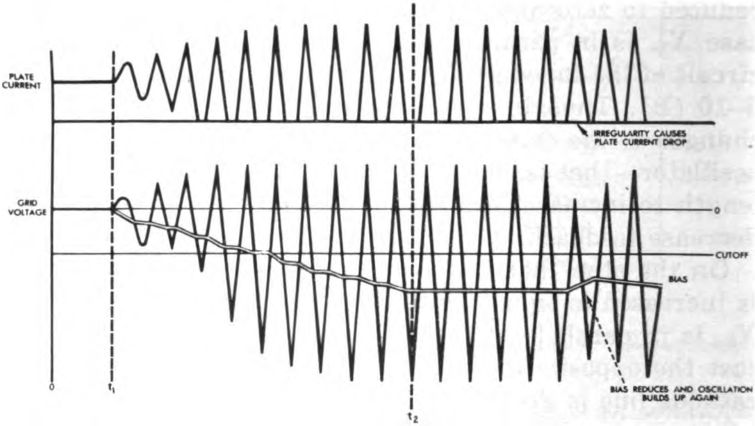


Figure 6–13.—Plate current and grid voltage vs. time, showing buildup of oscillations.

tude of oscillation is limited and stabilized. Although you are probably already familiar with grid leak bias, it is discussed briefly again to introduce the subject of self-pulsing. In this connection notice the graphs of plate current and grid voltage versus time in figure 6–13. These waveshapes are the same for all oscillators which use grid leak bias. At zero time, plate voltage is applied but the grids are grounded so that a steady plate current flows. At time t_1 the ground is removed, and any random change in plate current will cause a change in voltage in the plate tank circuit, a portion of which is fed back to the grid circuit. Since this feedback is in the right phase and amplitude, the oscillations will grow in amplitude. Now each time the grid goes positive, grid current flows and charges the grid leak capacitor more negatively so that on each successive oscillation the average grid voltage (the bias) goes more negative. This continues until a state of balance is reached where just enough grid current is drawn to replace the charge which leaks off through the grid leak resistor during the remainder of the cycle. In addition as the average grid voltage becomes more negative, the plate current flows for a shorter portion of each cycle until the flow is just sufficient to replace the energy lost in the tank circuit and to supply feedback to the grid circuit.

With perfect circuit elements and smooth electron emission, the circuit would oscillate indefinitely at this amplitude. But minute irregularities occur in the movement of electrons through resistors and conductors; also the emission from a cathode is always nonuniform. The nonuniformity is slight, but enough to affect the equilibrium in the oscillator circuit. Any slight decrease in current decreases the feedback, and if the time constant of the R–C circuit is not extremely long, the bias is reduced slightly by discharge of the capacitor. The decrease in bias allows a little more plate current, which reestablishes the equilibrium at the previous level.

In a circuit with a long R–C, however, the circuit starts oscillating, then settles down to a balanced condition as previously described, but any slight decrease in plate current now will not be compensated for by a lowered bias as shown in figure 6–14. The capacitor discharges too slowly. The resulting continued high bias

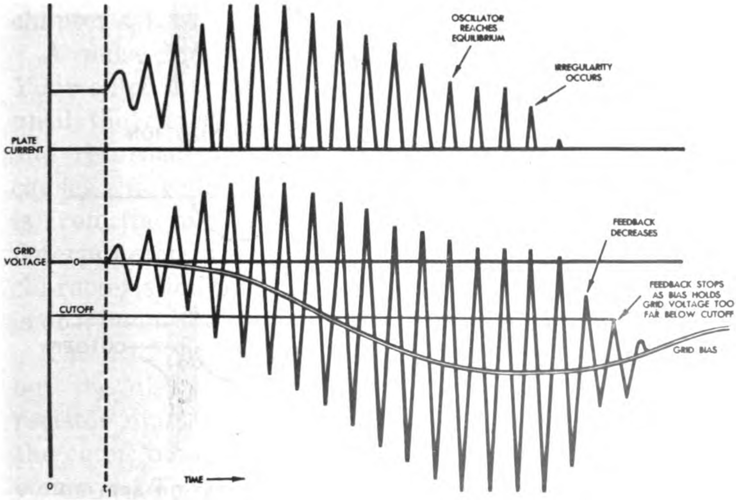
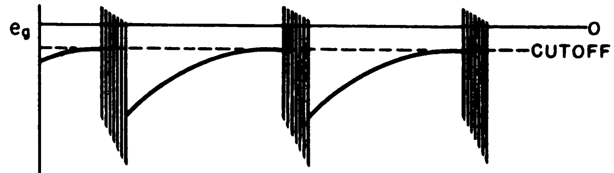

OSCILLATOR REACHES EQUILIBRIUM
IRREGULARITY OCCURS
PLATE CURRENT
GRID VOLTAGE
0
CUTOFF
FEEDBACK DECREASES
FEEDBACK STOPS AS BIAS HOLDS GRID VOLTAGE TOO FAR BELOW CUTOFF
GRID BIAS
TIME
0
t1

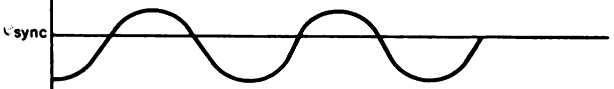
Figure 6–14.—Effect of long R–C.

prevents the feedback from being restored to the proper level and after a few cycles, the oscillation dies out completely. The capacitor slowly discharges and when its voltage drops below the cutoff voltage, the current again flows and the process repeats. For continuous operation, the R–C of the self-biasing circuit must be short enough to allow automatic adjustment of the bias for changing conditions in the circuit. To make the circuit oscillate briefly, then stop, use a long R–C. You can see that the time the tube remains cut off depends upon both the grid capacitor and grid resistance. By properly choosing these values, you can have the oscillator operate for 1 or 2 microseconds and then cease operation for 500 to 1,000 microseconds.

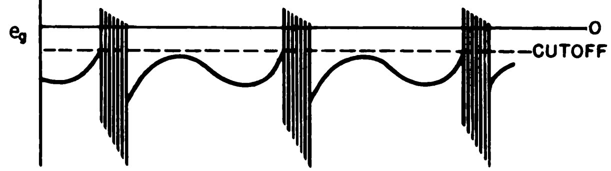
Grid voltage variations are shown in figure 6–15. It is not possible to show the actual time relations since the OFF period is several hundred times as long as the ON period.



(A) GRID VOLTAGE WITH NO SYNCHRONIZATION



(B) SYNCHRONIZING SINE WAVE



(C) GRID VOLTAGE WITH SYNCHRONIZATION APPLIED

Figure 6–15.—Grid waveshape with and without synchronization.

In equipment which uses a self-pulsing oscillator it may be necessary to synchronize the oscillator with other circuits. One method of doing this is by applying a synchronizing sine wave to the grid. The results are shown in figure 6–15 (B) and (C). Without synchronization, the grid rises exponentially and crosses the cutoff line at a small angle which means that the actual time that oscillation begins may vary somewhat. Notice in (C) that the point at which the grid voltage crosses the cutoff bias line is much more definite. Other ways of controlling the PRF are to feed in a positive-going pulse when the grid voltage approaches cutoff or to return the grid to a positive voltage rather than to ground.

In order to make the duration of the pulse more definite, some self-pulsed oscillators use an artificial transmission line instead of the grid leak capacitor. This has the added advantage of making the operating conditions more nearly constant throughout the pulse. There are two places where the pulse line can be located. One is to put it in the grid circuit, as shown in figure 6–16, the other, in the cathode circuit. (The construction and operation of artificial transmission lines are explained in chapter 4.)

A pulse line behaves like a long transmission line. Your chief interest is that it acts like a resistive circuit until the charging pulse has traveled the length of it and returned, at which time it is fully charged. This causes the voltage to change in two steps, the first step is from the bias E (figs. 6–16 and 6–17) to the voltage determined by the product of the grid current times the characteristic impedance of the line, and the second step is of a magnitude equal to the first.

The PRF is controlled by the synchronizing pulses but the discharge of the pulse line through the grid resistor must be such that the grid voltage approaches the cutoff point by the time the next synchronizing pulse comes in. The discharge of the line is in steps, as shown in figure 6–17, which is the diagram of the grid wave-

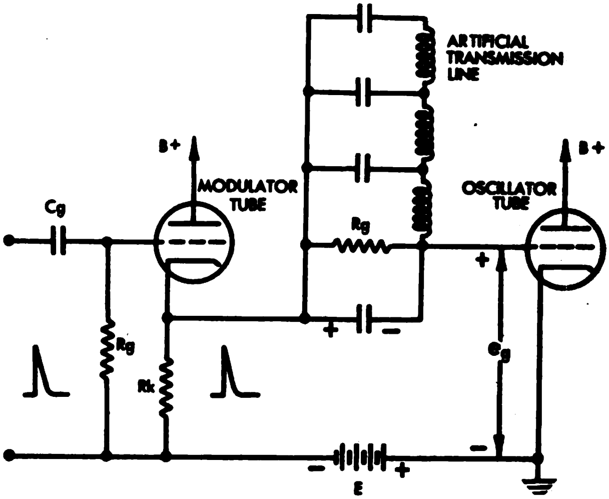


Figure 6–16.—Oscillator circuit with pulse line in grid circuit.

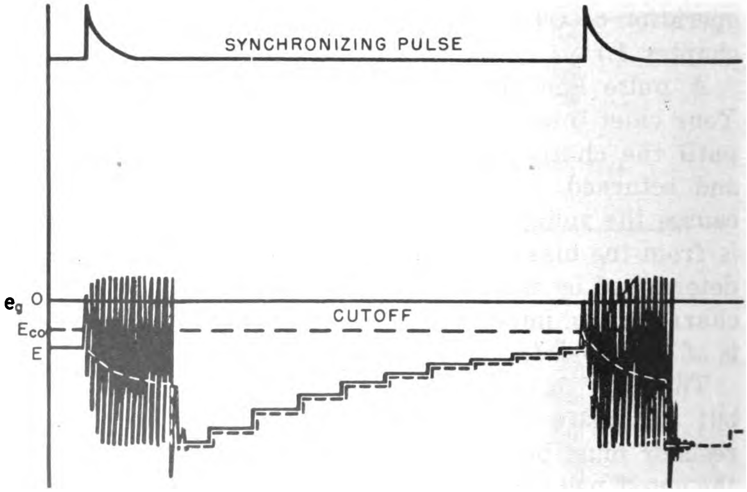


Figure 6–17.—Grid waveshape of a line-controlled self-pulsed oscillator.

shape of a line-controlled self-pulsed oscillator. The magnitude of the discharge voltage is determined by the relative values of resistance of R_g and the characteristic impedance of the line. R_g is usually much larger, hence the steps are small. These steps follow the general path of the exponential discharge curve with time constant R_gC_o where C_o is the capacitance of the pulse line.

SEPARATE MODULATORS.—As previously mentioned, some triode UHF oscillators are not self-pulsed, but employ separate modulators. These modulators are very much like those used with magnetron oscillators, which are discussed later in this chapter. Here merely note that the oscillator may be plate modulated, grid modulated, or cathode modulated.

Figure 6–18 shows a circuit for plate modulation in which short positive rectangular pulses are applied to the grid of the modulator tube resulting in a negative-going pulse at the plate. The transformer is connected so that plate voltage is applied to the oscillator tube or tubes in short rectangular pulses. RF components for the oscillator are not shown in the diagram. The transformer is specially constructed so that it passes the rectangular pulse without undue distortion. Since the purpose of the transformer is polarity inversion, the question arises as to why negative-going pulses are not applied to the grid of the modulator tube to cut it off and obtain a positive-going pulse at the plate for use in modulating the oscillator. This would require that the modulator tube, or tubes, conduct heavily during all the cycle except during the pulse and would result in excessive waste of power. The grid capacitance C_g of the oscillator is quite small, usually consisting of only the capacitance to ground of the grid-tuned circuit. If made too large, C_g might cause the oscillator to be self-pulsed and cut itself off before the modulator pulse is over. Hence C_g and R_g have values that would cause continuous wave operation if plate voltage were applied all the time. The modulator pulse then is most effective in controlling

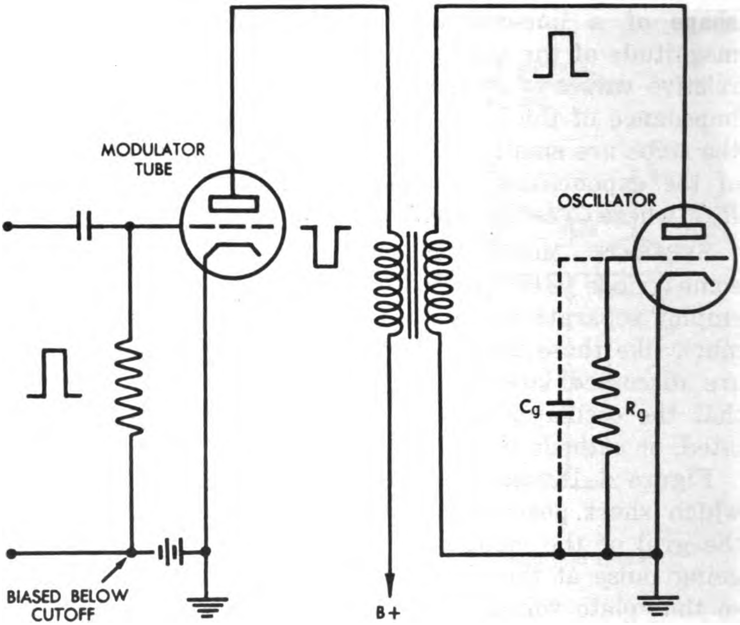


Figure 6–18.—Plate-modulated oscillator.

the output pulse. Oscillations must build up rapidly so that the leading edge of the output pulse will be as steep as possible. The plate voltage of the oscillator must remain constant throughout the pulse.

A cathode-pulsed (modulated) oscillator does away with the necessity of inverting the pulse, hence no transformer is used. (Refer to fig. 6–19.) The charging diode serves to recharge the coupling capacitor *C* between pulses. The capacitor must have quite a large capacitance so that it will not discharge appreciably during the pulse and thereby lower the plate-to-cathode voltage.

Grid modulation requires less power from the modulator for operation than either of the other two types of modulation. In grid modulation a positive-going rectangular pulse is applied to the grid and raises the grid voltage well above the cutoff point so that oscillation starts. Oscillation ends whenever the grid voltage drops

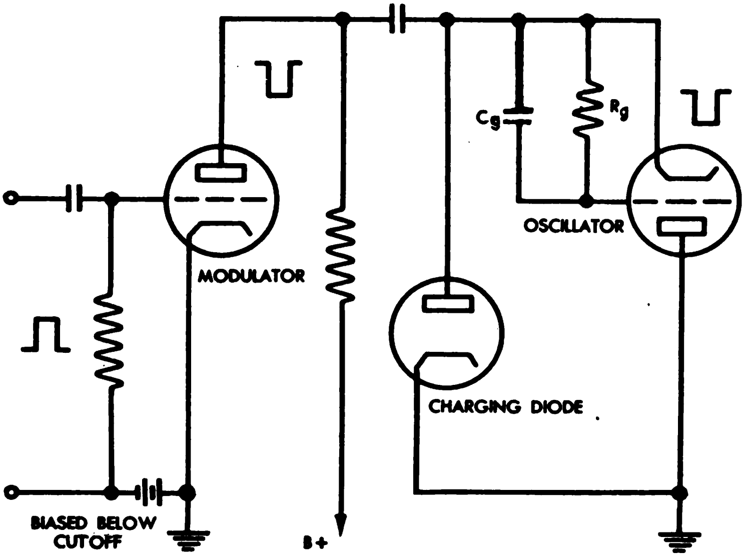


Figure 6–19.—Cathode-modulated oscillator.

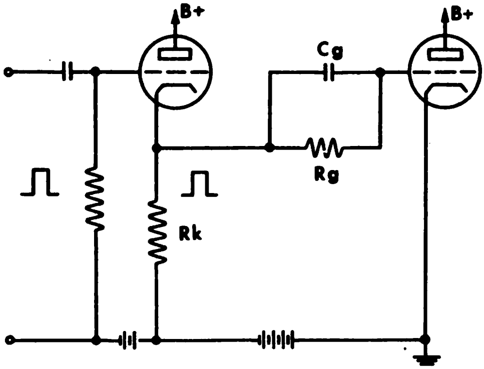


Figure 6–20.—Grid-modulated oscillator.

to the normal bias. Usually the arrangement for this purpose is the one in which oscillation is started by a positive trigger pulse and ended by a self-pulsing grid bias voltage. C_g in the grid-pulsed oscillator shown in figure 6–20 is a small capacitance and is used to couple the pulse from the cathode of the modulator tube to the grid of the oscillator to insure a rapid rise of voltage at the grid and cause oscillations to build up quickly.

THE MAGNETRON OSCILLATOR AS A TRANSMITTER

A magnetron is used as the transmitting tube in almost all pulsed transmitters that operate at frequencies above 1,000 mc. The fact that transit time has little influence on the magnetron, whereas it is a limiting factor in the operation of conventional tubes, makes this type tube desirable for UHF pulsed operation. Some triodes may be used in transit-time oscillator circuits at radar frequencies, but their use is limited because of lower power output. There are some velocity modulated tubes, such as the klystron, which operate in the radar frequency range and are capable of greater power output than the triodes, but they too are not desirable since they are not particularly adaptive to high-power pulsed equipment used in present-day aircraft.

Magnetrons are capable of being operated at more than 50 percent efficiency and can produce pulsed power outputs at frequencies as high as or possibly higher than 25,000 mc. The power output of a magnetron may exceed 100 kilowatts at 10,000 mc. and 5 megawatts at 3,000 mc.

The magnetron, which is a self-excited oscillator, possesses rather poor frequency stability as compared with the more stable lower frequency oscillators (electron-coupled or crystal-controlled). However, by the use of AFC and proper design the magnetron provides adequate stability. The magnetron is a type of diode in which a magnetic field is set up perpendicular to the electric field existing between the cathode and the plate. The magnetic field is provided by an electromagnet in some instances but in most cases by a permanent magnet. Figure 6–21 shows a cutaway view of a magnetron tube. The tube consists of a cylindrical plate with a cathode placed coaxial with it. The tuned circuits in which oscillations take place are the resonant cavities in the plate.

The magnetron tube is a delicate instrument and must

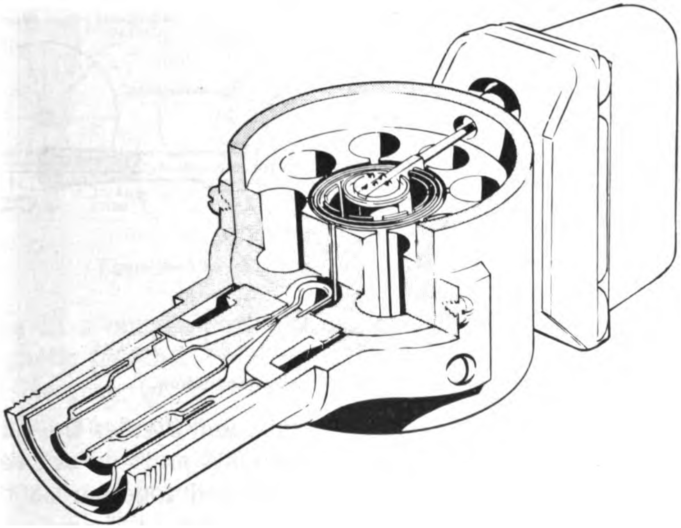


Figure 6–21.—Cutaway view of a magnetron.

be treated as such if satisfactory operation is obtained. Handle the tube carefully, for should it be dropped or should the tube elements be moved there is the possibility of a plate-to-cathode short or at least a change in spacing of these elements. Either of these will cause the tube to operate improperly. The magnets, which are made of a special alloy called Alnico, are unusually strong. The field strength required is dependent upon anode potential and operating frequency. The magnets should not be struck with metallic substances or any hard material for this will cause them to lose magnetism.

A MULTICAVITY TYPE MAGNETRON.—Various types of magnetrons have been designed, but at the present time the most used type is the resonant-cavity (traveling-wave) magnetron. This is the only type that will be discussed in this chapter. A magnetron of this type is shown in figure 6–22. The plate structure is a solid block of copper in which the resonant cavities are cylindrical holes. (Figure 6–22 is an illustration of the magnetron shown in figure 6–21.) A narrow slot opens each cavity

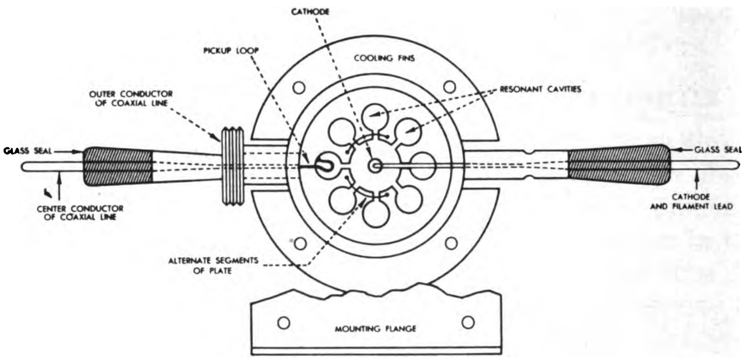

CATHODE
PICKUP LOOP
COOLING FINS
OUTER CONDUCTOR OF COAXIAL LINE
RESONANT CAVITIES
GLASS SEAL
GLASS SEAL
CENTER CONDUCTOR OF COAXIAL LINE
CATHODE AND FILAMENT LEAD
ALTERNATE SEGMENTS OF PLATE
MOUNTING FLANGE

Figure 6–22.—Multicavity magnetron.

into the central portion of the tube and divides the inner plate structure into as many segments as there are cavities. Alternate segments are strapped together so that the cavities are placed in parallel so far as the output is concerned. This makes it possible to take the output from a pickup loop placed inside any one of the cavities. Since the outer conductor of the output coaxial line is at ground potential and connected to the shell of the magnetron, which is a part of the plate structure, there cannot be a high positive d-c voltage applied to the plate. The same result is realized whenever negative voltage is applied to the cathode, since it is equivalent to driving the plate positive.

Operation of the Magnetron Oscillator

The theory of operation of the magnetron is based on the motion of electrons in combined electric and magnetic fields.

EFFECT OF MAGNETIC FIELD ON ELECTRON FLOW.—When no magnetic field exists, heating the cathode results in a uniform and direct electron movement from the cathode to the plate surrounding it (fig. 6–23 (A)). However, as the magnetic field surrounding the tube is continually increased, a single electron is affected as shown in figure 6–23 (B), (C), and (D). In (B) the magnetic field has increased to a point where the electron proceeds to the

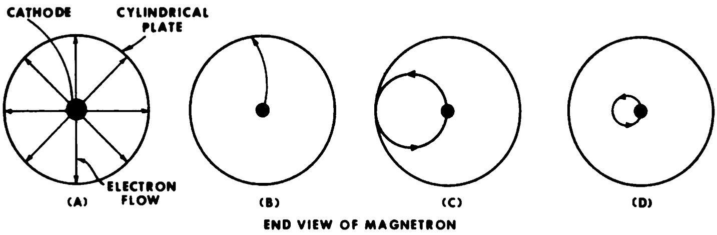


Figure 6–23.—Effect of magnetic field on electron.

plate in a curved rather than direct path. In (C) the magnetic field has reached a value great enough to cause the electron to just miss the plate and return to the cathode in a circular orbit. This value is the critical value of the field strength. In (D) the value of the field strength has been increased to a point beyond the critical value and the electron is made to travel to the cathode in a circular path of smaller diameter. Thus the size of the circle or the path that the electron follows around the cathode for a fixed cathod-to-anode potential will depend upon the strength of the magnetic field. Since these paths are determined by both magnetic and electric fields they may be altered by changing the strength of either. The strength of these fields is critical for they determine the paths that the electrons follow and this, in turn, determines the stability of oscillation and tube efficiency. Under ideal conditions the electron paths are such that when curving past the slots of the resonant cavities energy is released to the RF field.

PRINCIPLE OF OPERATION.—The principle of operation of a magnetron is expressed in the process by which an electron acquires energy from the d-c field (existing between the anode and cathode) and releases it to the RF field to build up and sustain oscillations. (This happens when the d-c field accelerates the electron and causes it to acquire energy which is proportional to electron velocity.)

The cavity resonators are essentially a number of resonant circuits which are all coupled together. The RF fields of these resonators may or may not be in phase. This phase relationship is determined by the amount of coupling between the cavities. Desired coupling may be acquired by connecting selected segments together. This is known as STRAPPING. By referring to figures 6–21 and 6–22 you can see an application of strapping.

Magnetrons may operate at different modes, which is to say that they may resonate at different frequencies. These frequencies are determined by the velocity at which the electrons whirl about the anode. A mode of operation that produces the greatest power output and efficiency is the π mode. The resonator is said to operate in the π mode when the phase difference of 180 degrees exists between successive cavities.

The electric component of the RF field at a given instant may be represented by dotted lines as shown in figure 6–24. In this figure, alternate segments are connected together or strapped to maintain a phase difference of 180 degrees between segments (π mode). The figure can be redrawn as a plane-electrode magnetron with field distribution as shown in figure 6–25. If motion of

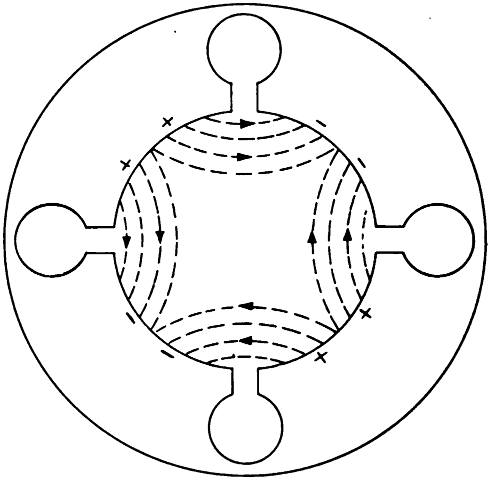


Figure 6–24.—Electric field distribution in a four-cavity magnetron.

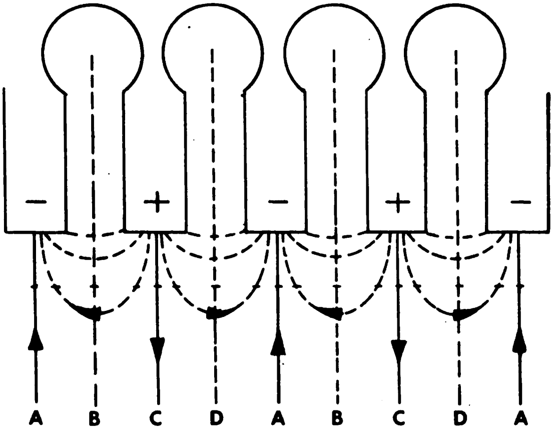


Figure 6–25.—Force lines in a plane-electrode magnetron.

the electrons in figure 6–25 is considered, an electron at point *B* is accelerated to the right, one at point *C* is accelerated vertically, and one at point *D* is accelerated to the left. This action produces bunching so that bunches of electrons pass the gaps the instant the field causes maximum deceleration. The electrons yield energy to the RF field, and induce strong oscillations in the resonant cavity of which the gap is a part.

Since the RF field changes in synchronism with the electron motion, the field lines of figure 6–25 may be considered as a traveling wave moving to the right in synchronism with the electrons. The bunching mechanism produces a spoke-shaped cloud of electrons that moves around the cathode in synchronism with the anode wave, as shown in figure 6–26.

As electrons are emitted from the cathode, they combine with the spoke-shaped field and simultaneously spiral towards the anode. Exceptions to this are the electrons that are emitted from the cathode at such time that they absorb energy from the RF fields. These will be quickly returned to the cathode. When electrons approach the opening of the resonate cavity they will give energy to the cavity providing the RF voltage across the opening

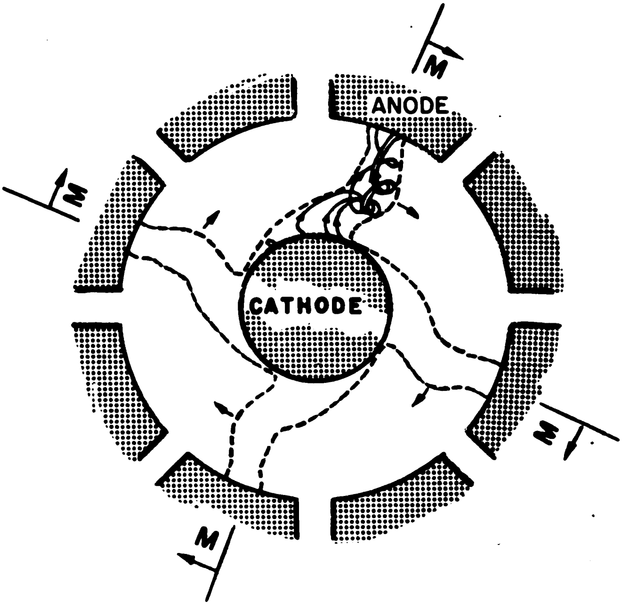


Figure 6–26.—Orbits of electrons of different phases in a cavity magnetron.

is a decelerating field. (This is illustrated in figure 6–27.) Electrons on the RF negative side of the anode will be repelled by the approaching electrons. This provides the necessary "kick" to the oscillatory action of the cavity and thereby oscillation is sustained. As the electrons pass each cavity, they are slowed down when some of their energy is given up. After a few revolutions their velocity has decreased to the extent that they no longer continue in orbits but are attracted to the anode.

MODES OF OSCILLATION OF THE MAGNETRON.—There are various modes of operation of a magnetron and they are identified by the number *n*, which specifies the number of cycles of variation of the field quantities that occur around the circumference of a circle surrounding the cathode. For each mode of oscillation of a magnetron, like those of other cavity resonators, there is a particular pattern of field variations. In the *n* equals zero mode,

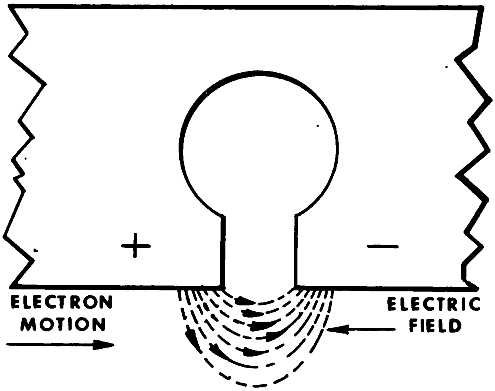


Figure 6–27.—An electron approaching the slot of a cavity at a time when it will give up energy in the cavity's RF field.

since the fields do not enter the anode cavities, the frequency of the mode is not controlled by the cavities; thus, this is an undesirable mode of operation. With the exception of the *n* equals zero and π modes, all modes are degenerative. The degenerative modes are in reality a combination of two modes that have the same resonant frequency. Due to unavoidable unsymmetrical construction of the anode segments the degenerative modes separate into two modes of slightly different frequencies.

Since a slight variation may cause a shift from one of these modes of operation to the other with a resultant change in output frequency, these modes are highly undesirable. Since the π mode of operation (previously described) is not degenerative, it is the desired mode of operation.

It is possible in an unstrapped magnetron to have an undesirable mode whose frequency may differ from the π mode frequency by less than 2 percent. The undesired mode will cause interference with the π mode of operation with such a small separation of frequency. With a strapped magnetron it is possible to obtain as much as a 10 percent separation and thus eliminate the interference.

Additional Features of Magnetrons

The power output of a magnetron is obtained either by a coupling loop, as shown in figure 6–22, or by a transformer section of waveguide inserted into a cavity, as shown in figure 6–28. The filament leads are brought in through the glass enclosure as shown in figure 6–21.

Cathodes of magnetrons usually are oxide-coated, indirectly heated surfaces, as shown in figure 6–29. The pulse currents developed in magnetrons are larger than can be produced by thermionic emission alone. The large

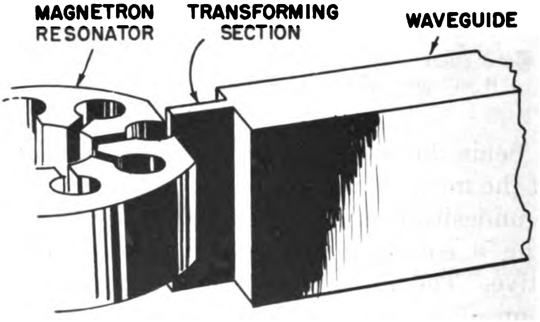


Figure 6–28.—Waveguide output section.

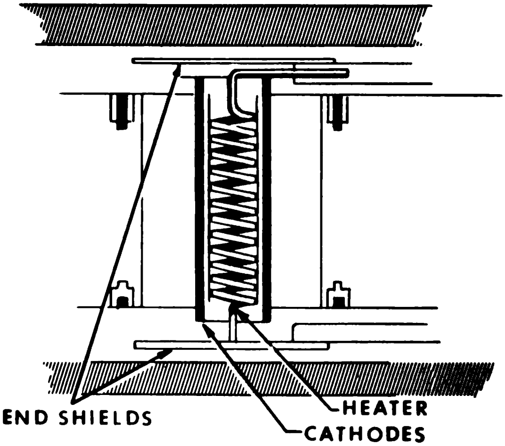


Figure 6–29.—Oxide cathode of a magnetron.

number of electrons are developed by secondary emission from the cathode as the cathode is back-bombarded by high-velocity electrons which are returned by the magnetic field. Metallic cathodes with good secondary-emission characteristics have been used satisfactorily. The cathodes are designed with end shields to prevent the escape of electrons from the interaction space between the cathode and anode block. The design of anode blocks depends on the frequency and power output.

Another type of anode block, other than that shown in figure 6–21, is one that can be used unstrapped and is called the RISING SUN because of its physical shape. This magnetron, in π mode operation, requires no strapping because its irregular shape contributes to stability.

Magnetrons are difficult to tune. Since the resonant circuits are within the vacuum, they are inaccessible and mechanical tuning is slow. Electronic tuning can be used over a narrow frequency range and permits frequency modulation. Mechanical tuning is done by movement of a mechanical or magnetic device. Ranges of about plus or minus 10 percent of the mean frequency can be obtained by mechanical-tuning methods, two of which are shown in figure 6–30. In (A), movable tuning pins are inserted in each of the resonator cavities in order to vary the cavity inductance. These pins are made of a nonmagnetic material. When these tuning pins are inserted into the cavity they reduce the space available for the magnetic flux and thus the inductance is reduced and the frequency is increased. In (B), tuning is accomplished by varying the capacitance of the resonator. This is accomplished by moving a U-shaped ring in or out of grooves cut into the anode block. When the ring moves into the grooves the frequency of the magnetron is increased.

An application of magnetron frequency control as illustrated in (B) of figure 6–30 is used to frequency modulate the magnetron used in the aircraft altimeter, AN/APN–22. In this equipment the magnetron is modu-

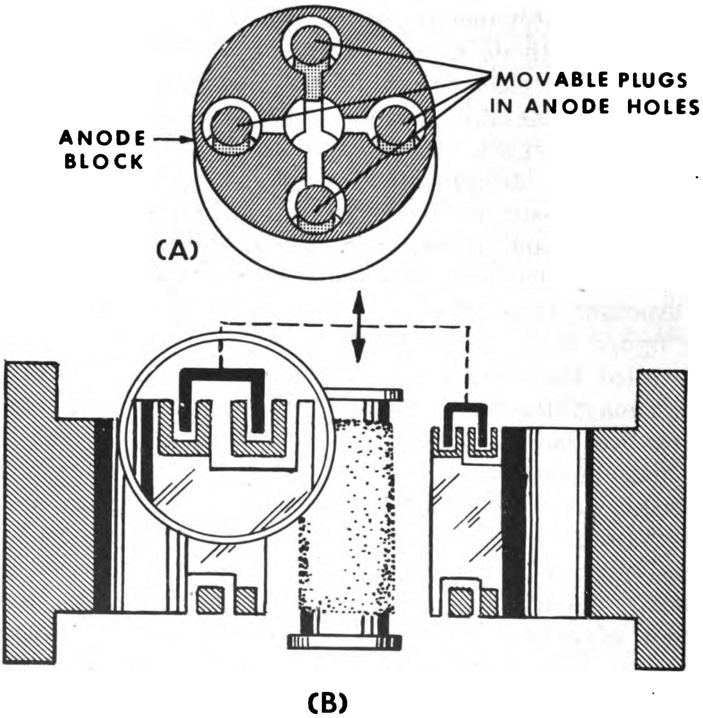


Figure 6–30.—Two types of tuning methods.

lated by a vibrating reed which varies the capacity across the strap on the anode cavity.

Since much of the heating of the cathode is obtained by the bombardment of the cathode by the electrons returning to it, the filament voltage is usually reduced when the high voltage is applied. This reduction is accomplished by adding a tapped filament transformer or a series limiting resistor in the circuit. In some equipments the filament voltage is removed completely.

Depending upon the type of magnetron and the type of equipment, it is sometimes necessary to season a new magnetron since it may not withstand the immediate application of full-power plate voltage. This is accomplished by subjecting the magnetron for a period of a few hours to filament voltage only. After this warmup period the plate voltage is applied (at low-power settings if possible) and a check is made to determine if the magnetron is arcing. Should arcing be detected, it may be necessary to remove the plate voltage and again subject the magnetron to an additional period of seasoning. The *Handbook of Service Instructions* will generally contain information in connection with this phase of maintenance.

In airborne equipments that use a magnetron the magnetic field is provided by a permanent magnet. Some magnetrons, called PACKAGED MAGNETRONS, have built-in magnets; thus, whenever a magnetron is replaced the magnet must also be replaced. The packaged magnetron is finding wide use at the present time.

Theoretically, if the magnetic field of the magnetron is reversed, there should be no change in magnetron operation. This may not always be true since the construction of the magnetron may not be perfectly symmetrical. Thus, after removing a magnet, it is important that it be replaced in its original position. Generally the polarity of a magnet is indicated and this may be used as an aid in maintenance. The magnet should never be removed when the magnetron is operating since this will cause the electrons to strike the anode at high velocity and cause excessive heating. With no magnetic field there is nothing to limit magnetron current.

Modulating the Magnetron

The modulator for the magnetron serves a double purpose—it determines the waveform of the output pulse, and it stores energy between pulses and releases it through the magnetron during the pulse.

The basic circuit for a magnetron modulator is shown in figure 6–31. In this circuit the storage element is charged through the charging impedance Z_{ch} and the load. It is discharged through the load when the switch is closed. On examining this circuit, you can see that closing the switch also completes a circuit containing

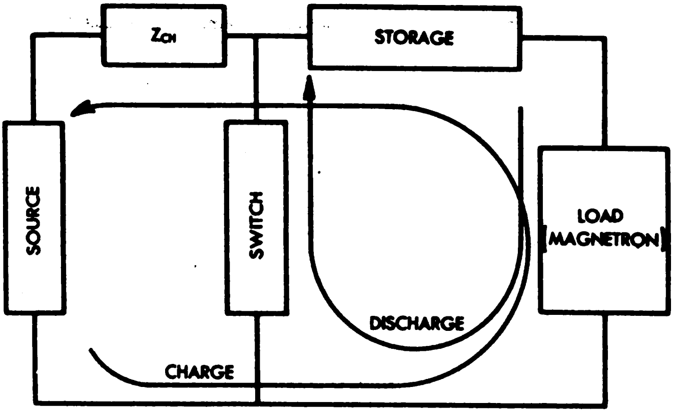


Figure 6—31.—Basic circuit for storage and release of energy in a modulator.

the power source and Z_{ch}. This has little consequence, however, since the switch is closed for only a microsecond or so at a time and Z_{ch} is large enough to prevent any damage. It is well to become familiar with the components of some of the modulator circuit elements which are shown as blocks in this circuit. In the modulator, the source from which the storage element is charged may be either a d-c voltage or an a-c voltage. The magnetron requires a voltage of 10 to 15 kilovolts or higher for proper operation. In some types of equipment a transformer is used between the storage element and the magnetron to step up the voltage while in many others a direct connection is made between the storage element and the magnetron. Actual charging circuits are discussed in a later section dealing with line-pulsing modulators. In these circuits the charging impedance may be a high resistance or a high inductance, or both. Its purpose is to control the charging time for the storage element and to prevent short circuiting of the source during the pulse.

SWITCHING DEVICES.—Switches for controlling the discharge time of pulse-forming networks may be of various types. Three types that will be discussed in this chapter are the ROTARY SPARK GAP, GAS DIODE, and THYRATRONS.

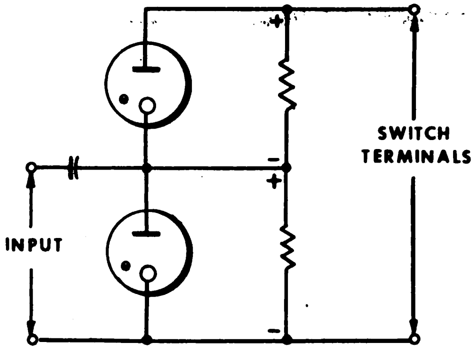


Figure 6–32.—Gas diode switch.

Rotary spark gaps are generally not used in aviation equipment because they are instable, must be enclosed in a pressurized chamber, and different repetition rates are difficult to obtain.

Gas diodes are used in some high-powered modulators. Figure 6–32 illustrates a gas diode type switch in which two cold cathode gas diodes are used. Some equipments use more than two of this type diode. These tubes are connected in series and an equal resistance is connected in parallel with each so that the high voltage will divide equally across each tube. This division prevents the voltage across any tube from becoming high enough to cause the tube to conduct.

As illustrated in figure 6–32, the trigger pulse is applied across one diode and causes it to conduct. This conduction causes the triggered diode to appear as a closed switch which connects the full high voltage across the other diode and causes it to conduct. The conduction of this tube closes the network circuit. This circuit remains closed until the storage network discharges to the point where the diodes de-ionize. At this time the storage network begins to recharge.

The thyratron is the more common switching device utilized in modulators. It is operated at cutoff until a positive trigger pulse is applied to the grid. The tube then conducts until the network circuit becomes discharged to the point that the thyratron de-ionizes and the grid regains control. The most common thyratron of this type is the HYDROGEN THYRATRON. While the gas diodes require an extremely high trigger voltage, the hydrogen thyratron requires a much lower voltage, approximately 30 to 100 volts. The normal life of this type of thyratron is expected to be about 900 hours; tube life is influenced by repetition rate and not pulse widths, providing the widths do not exceed 5 microseconds. The two advantages of the thyratron switching system are the accuracy of the firing time and the fact that the thyratron is not affected by the impedance of the trigger source.

PULSE-FORMING NETWORKS.—The pulse-forming network consists of a series of coils and capacitors in an arrangement like that shown in figure 6–33. There are other arrangements for the elements in the PFN, but the arrangement shown is the most common. Most pulse-forming networks have a CHARGING CHOKE in series with the input. The choke is series resonant with the line and isolates the power supply from the switching device during discharge time. The LC constants in the network are carefully selected to provide a pulse of the proper shape and duration. The coils and capacitors must be rated to withstand the high voltages and currents they encounter in building up the high-powered pulse.

The PFN and charging choke are designed so that their combined impedance matches the output load as nearly as possible when the network is switched. The output line therefore appears as an infinitely long transmission line when the switch is closed. This reduces reflections in the charging circuit during the discharge interval. During the charging time, however, when the switch is open, the network is resonant. It can be shown that when a voltage is suddenly applied to an inductance and capacitance in series, the voltage at the resonant frequency of the circuit oscillates in such a manner as

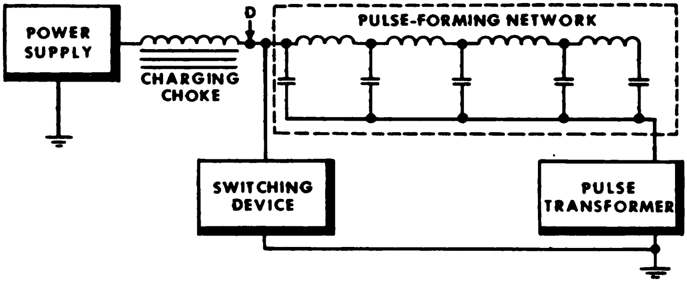


Figure 6–33.—Schematic diagram of a typical pulse-forming network and charging choke.

to give a sine wave, the amplitude of which is almost twice the applied voltage.

For maximum voltage production the pulse recurrence frequency should be equal to twice the resonant frequency of the charging circuit. To permit operation at more than one repetition rate, a charging diode must be connected in series with the charging choke. This diode would be connected at point *D* in figure 6–33. Figure 6–36 shows a multi-PRF modulator using a charging diode. The pulse-forming network is thus prevented from discharging after the maximum oscillation is reached. The diode also causes some additional damping in the circuit, so that the maximum value is slightly less than the original maximum value of almost twice the value of the d-c supply.

Because the diode maintains the voltage at practically twice the d-c supply for a considerable time the interval between times of closing the switching device can have any value greater than a half cycle of the charging oscillation.

The output voltage pulse from a d-c resonant charging network is about half the voltage to which the network is charged because the line impedance and load impedance are equal. Since the network is charged to nearly twice the supply voltage and the output pulse voltage is half the network charge, the usable voltage in the output pulse is about 0.9 of the supply voltage.

This type of pulse network is usually sealed in a metal box filled with oil which acts as an insulator. For that reason it is sometimes called a POTTED NETWORK. It is impossible to replace components in the line. Some lines using only a few elements are not potted. These do not generally give as nearly a rectangular output pulse, due to the smaller number of sections.

THE PULSE TRANSFORMER.—In the circuits discussed thus far the load can be a magnetron and its associated circuits or the primary of a pulse transformer which has its secondary connected to the magnetron. The use of a high-voltage step-up transformer to furnish voltage for the magnetron makes possible the use of lower voltages in the pulse-forming network and thus simplifies the insulating problem. An advantage in using a transformer with two secondaries (bifilar windings), as shown in figure 6–34, is that it avoids having high voltage in the magnetron filament transformer. The bifilar windings have the same voltage induced into each since they are wound side by side. This winding makes it possible to connect a filament transformer, without high-voltage insulation, to the heater of the magnetron and to couple the high-voltage pulse between the cathode and plate.

A third need for the high-voltage pulse transformer is matching impedances, particularly where the radar antenna is remote from the set itself. Since the use of long transmission lines between the magnetron and the antenna make the magnetron's operation somewhat erratic, it is desirable to have the magnetron near the antenna. This necessitates the use of long lines to carry the modulator pulse to the magnetron. In order to be able to use 50-ohm line for this purpose and still obtain an impedance match with the impedance of the magnetron (approximately 1,000 ohms), it is necessary to use a 4½ to 1 step-up transformer at the magnetron end.

The pulse transformer must be specially designed because there are very high frequency components present in a rectangular pulse of such short duration. The core

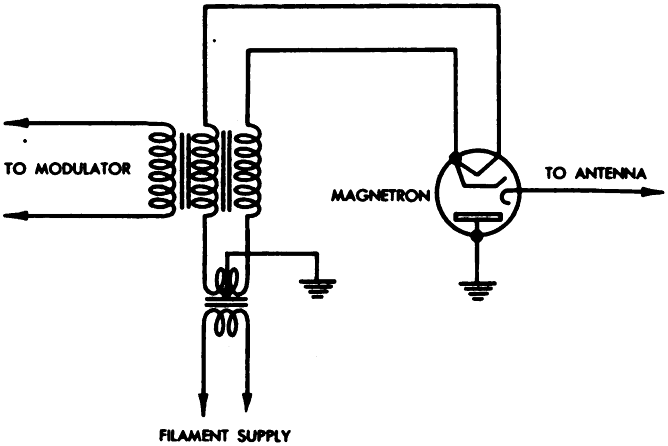


Figure 6–34.—A transformer with two secondaries makes possible the grounding of the filament transformer secondary.

is usually silicon steel and has laminations in the order of 0.003 inch in thickness. In general a good pulse transformer must have low leakage inductance, low interwinding capacitance, and high primary inductance.

Types of Modulators

There are two types of modulators—the line-pulsing modulator and the driver-hard-tube modulator. The line-pulsing modulator stores energy and forms pulses in the same circuit element. This element is usually the pulse forming network. The driver-hard-tube modulator forms the pulse in the driver; the pulse is then amplified and applied to the modulator. The hard-tube modulator has been replaced by the line-pulsed modulator. The reasons for this replacement are that the hard-tube modulator has lower efficiency, its circuits are more complex, a higher power supply voltage is required, and it is more sensitive to voltage changes.

The line-pulsed modulator is easier to maintain because of its less complex circuitry and, for a given amount of power output, it is more compact and light. Since it is the principally used modulator in aviation radar, it is the only type that will be discussed at great length.

Before discussing the line-pulsing modulator, it is well to consider briefly the various types of charging circuits and to note their advantages and limitation. To do this, study figure 6–35 which shows the various types of charging circuits.

In the d-c resistance charging circuit (A), the pulse line or storage capacitance is represented as C_{st}. It is charged through R_{ch} to the value of the d-c voltage. During the pulse one-half the d-c voltage is applied to the magnetron (assuming the line impedance is matched to that of the magnetron) for a period of time determined by the line components. This is only 50 percent efficient. Therefore, it is necessary that the d-c voltage be twice the value needed for the magnetron. The time constant $R_{ch}C_{st}$ must be large in comparison to the period of the repetition frequency. This insures that the line will fully charge between pulses but will not discharge appreciably through the source, and E_{ch} during the pulse.

In the d-c resonance charging circuit (B), resistor R_{ch} is replaced by an inductance L_{ch}. As C_{st} charges, the current through L_{ch} builds up a magnetic field. This field causes current to continue after C_{st} is charged to E_{bb}. The charge will reach a peak of approximately 1.9 times E_{bb} after which it starts decreasing. If the modulator is triggered at the instant of maximum charge, the voltage applied to the magnetron is about 95 percent of E_{bb}. This necessitates that the resonant frequency of L_{ch} and C_{st} be one-half the PRF of the radar. The firing time is very critical if advantage of the maximum voltage is to be obtained.

The difficulty encountered in the preceding circuit is overcome by adding a diode in series with the charging element. Circuit (C) prevents the line from discharging after reaching its peak. This means that the time of firing can occur any time after the peak is reached. Therefore, the frequency at which L_{ch} and C_{st} resonate is equal to or greater than one-half the PRF. The disadvantage of making the firing time less critical is that the

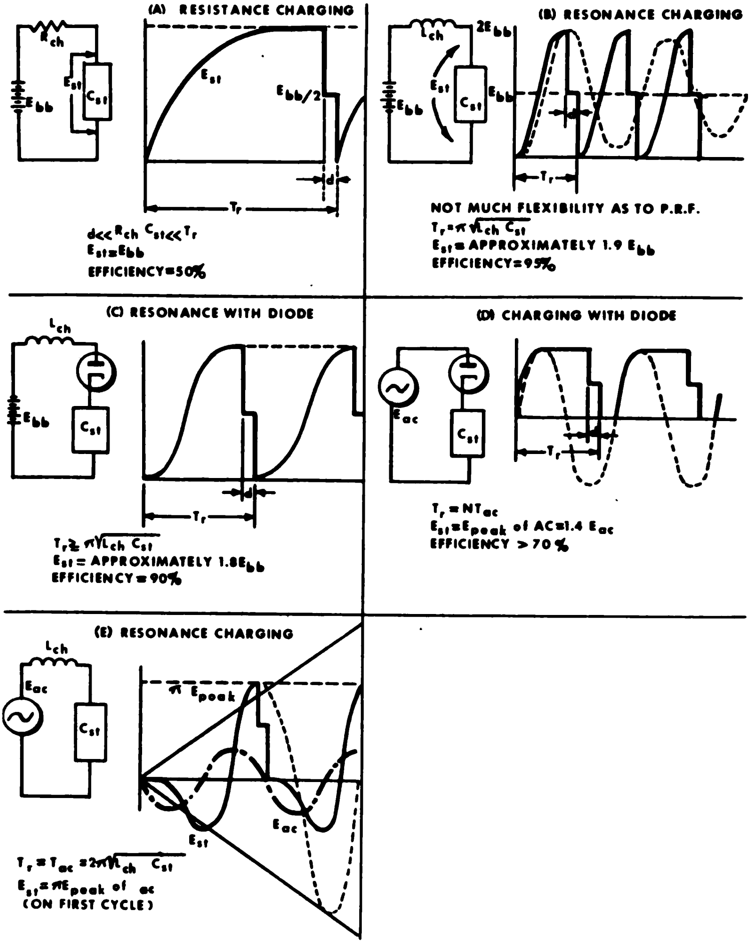


Figure 6–35.—Types of charging circuits.

voltage to which the pulse forming line charges is decreased by the drop across the diode to about 1.8 times E_{bb}. This makes the efficiency of this circuit about 90 percent.

Circuit (D) has the pulse line charging from an a-c source. In this circuit, the charging line is prevented from discharging by a diode in series with it. Further, in this circuit it is necessary that every trigger pulse come at a time when the diode plate is negative; otherwise the diode and the switch will short circuit the a-c source. Therefore,

$$T_r = NT_{ac}$$

where

T_r = time between trigger pulses

N = 1 or some other whole number

T_{ac} = time of a cycle of the a-c source voltage

The voltage to which the line charges is almost the peak value of the a-c wave so that the efficiency is about 70 percent.

In circuit (E), the diode is replaced by the inductance L_{ch}. This inductance, along with C_{st}, forms a resonant circuit which is resonant at the frequency of the a-c supply voltage. In this circuit the switch may be closed at any peak of oscillations. Usually it is closed at the first peak. At this time the voltage is approximately π times the peak value of the a-c voltage. In addition $T_r = T_{ac} = 2\pi \sqrt{L_{ch}C_{st}}$. If the switch is closed at the second peak, the voltage will be nearly twice as high. This type of charging gives voltages higher than the source voltage.

THE LINE-PULSING MODULATOR.—This discussion will serve a dual purpose—it will explain the theory of operation of a typical line-pulsing modulator and will also summarize the function of the various modulator components already explained in this chapter.

The modulator (fig. 6–36) is of the line-pulsing type. The pulse-forming network undergoes a charge-discharge cycle at a rate (PRF) which is established by the master-trigger generator and which is consistent with the pulse width in use and the maximum power capabilities of the transmitter. The modulator consists of artificial-line pulse-forming network Z401; hydrogen-thyratron switch tube *V*405; holdoff or charging diode *V*403; inverse-diode *V*404; resonant charging choke *L*401; and magnetron pulse transformer *T*404, the double-wound secondary of which supplies a pulse of power to the magnetron cathode circuit each time the line is discharged. The switching device which controls the discharge of the line is a hydrogen-thyratron tube controlled by the trigger from the radar's synchronizer.

When system power is applied, current to charge the pulse-forming network is drawn from the power supply through resonant charging choke *L*401 and charging diode *V*403. The flow of this current establishes a magnetic field in the resonant charging choke. Charging is rapid up to the point at which the pulse-forming network charge nearly equals the peak power supply voltage, approximately plus 2,500 volts d. c. The rate of charge drops sharply, and the magnetic field built up in the resonant charging choke collapses. By this self-inductance, a voltage approximately twice the power supply peak voltage is rapidly established. The tendency of the resonant charging choke to oscillate is halted abruptly at its first positive peak by the rectifying action of the charging diode and the pulse-forming network is held at this high potential. Discharge of the capacitors within the pulse-forming network is prevented by charging diode *V*403. Inverse-diode *V*404, and thyratron switch *V*405 are inoperative and hence appear as infinite impedances and do not provide a discharge path for the network.

The master-trigger generator in the synchronizer supplies a trigger voltage to the grid of the thyratron switch tube through the d-c blocking capacitor *C*402. When triggered the switch tube ionizes and provides a closed circuit for the discharge of the pulse-forming network. This discharge continues for a definite period of time at a constant rate, depending upon the constants of the pulse-forming network. A negative square-shaped pulse is impressed across the primary winding of the magne-

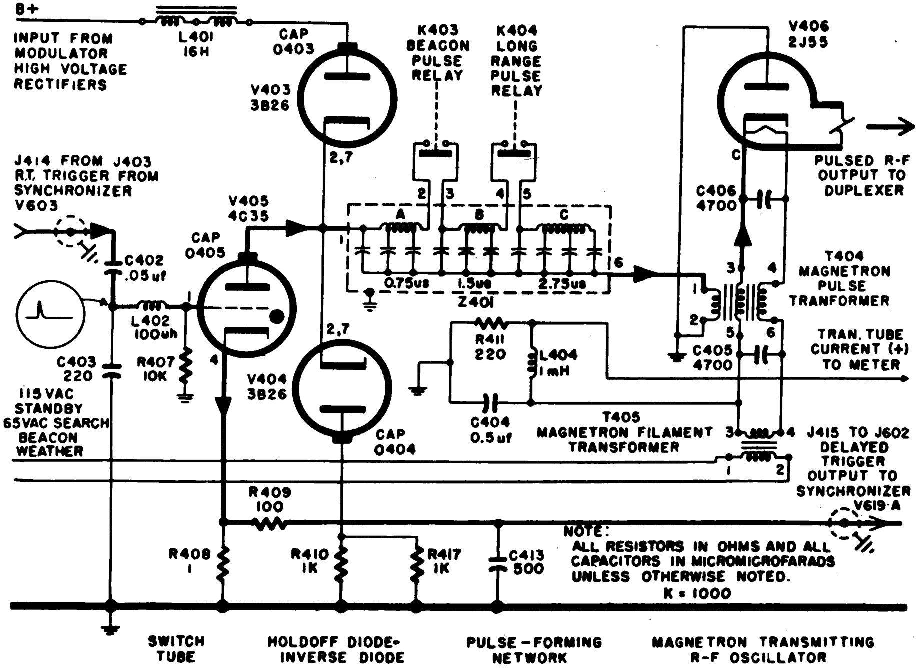

B+
INPUT FROM MODULATOR HIGH VOLTAGE RECTIFIERS
L401 16H
CAP 0403
V403 3B26
2,7
K403 BEACON PULSE RELAY
K404 LONG RANGE PULSE RELAY
V406 2J55
J414 FROM J403 R.T. TRIGGER FROM SYNCHRONIZER V603
C402 .05 uf
V405 4C35
CAP 0405
1
L402 100uh
R407 10K
4
C403 220
115 VAC STANDBY
65VAC SEARCH BEACON WEATHER
V404 3B26
2,7
CAP 0404
2
3
4
5
A
B
C
1
6
0.75us
1.5us
2.75us
Z401
R411 220
L404 1 mH
C404 0.5 uf
T405 MAGNETRON FILAMENT TRANSFORMER
C
C406 4700
PULSED R-F OUTPUT TO DUPLEXER
T404 MAGNETRON PULSE TRANFORMER
3
4
1
2
5
6
C405 4700
TRAN. TUBE CURRENT (+) TO METER
3
4
1
2
J415 TO J602 DELAYED TRIGGER OUTPUT TO SYNCHRONIZER V619-A
R409 100
R408 1
R410 1K
R417 1K
C413 500
NOTE: ALL RESISTORS IN OHMS AND ALL CAPACITORS IN MICROMICROFARADS UNLESS OTHERWISE NOTED.
K = 1000
SWITCH TUBE
HOLDOFF DIODE-INVERSE DIODE
PULSE-FORMING NETWORK
MAGNETRON TRANSMITTING R-F OSCILLATOR

Figure 6–36.—A typical line-pulsing modulator and magnetron circuit.

tron pulse transformer *T*404. The width of the transmitter pulse varies in a manner dependent upon the number of sections of the pulse line in use, as determined by the beacon pulse relay *K*403 and long-range pulse relay *K*404. Since the primary of transformer *T*404 is matched with the 50-ohm surge or characteristic impedance of the pulse-forming network or line, there is an equal division of voltage between these two units. Approximately 2,500 volts is impressed on the primary of *T*404. The turns ratio of *T*404 is 5 to 1, affording a voltage step-up of 1 to 5. Thus a negative pulse of approximately 12 kilovolts is placed on the cathode of the magnetron *V*406. The peak power of this pulse is in the order of 150 kilowatts. Output of the magnetron approximates 50 kilowatts due to losses in this tube and the associated circuits.

Gas bursts (arcing) occur occasionally within the magnetron. This condition causes momentary pulse-transformer impedance mismatching, and a small voltage of polarity opposite to that of the modulator pulse may appear across the pulse-forming network after the pulse. This small reverse voltage is in series with the power supply voltage at the start of the next charging cycle. Hence the total voltage available to produce the charging oscillation is increased, and the pulse-line voltage may build up to a value great enough to break down the thyratron. This condition is overcome by connecting inverse diode *V*404 in parallel with the thyratron. The diode bypasses the reversed charge of the pulse-forming network.

The pulse-forming network is composed of aircore inductors and mica capacitors. It is divided into three sections of artificial transmission line, all within the same container. Each part is a complete network. The connection of the three parts by relays *K*403 and *K*404 increases the width of the pulse. When only the first section *Z*401–A is in use, 0.75-microsecond pulses are generated. Addition of the second section increases the pulse period by 1.5 microseconds for a total of 2.25 microseconds.

With all three sections connected, pulses of 5.0-microseconds duration are generated.

During beacon operation only relay *K*403 is energized, thus the A and B sections of *Z*401 are connected in the circuit to form a 2.25-microsecond pulse. During long-range search operation both relays, *K*403 and *K*404, are energized thus all three sections of *Z*401 are connected in the circuit to provide a 5.0-microsecond pulse.

Resonant charging choke *L*401 not only doubles the pulse-system voltage before firing, but it effectively isolates the power supply from the rest of the modulator during the relatively short periods of pulse duration and prevents the power supply from shorting through the switch tube during de-ionization.

Resistor *R*407 functions as a grid leak for the thyratron switch tube. Capacitor *C*403 and choke *L*402 form a filter network to prevent the reflection of the switch tube transients into the synchronizer. Switch tube current, flowing through the cathode resistor *R*408, generates a trigger pulse which is used for synchronizing the range mark start-stop multivibrator located in the synchronizer. Resistor *R*409 and capacitor *C*413 improve the trigger shape by removing ripple at the top of the pulse.

The cathode of the magnetron is internally connected to one side of the filament line. Input to the magnetron is applied by the modulator through the magnetron pulse transformer *T*404. The design of the pulse transformer provides two secondaries of similar electrical characteristics, one in each leg of the magnetron filament supply. This bifilar secondary winding maintains an equal and in-phase pulse voltage on both sides of the filament and effectively isolates the high-voltage modulator pulse from the low-voltage magnetron filament transformer *T*405. Capacitors *C*406 and *C*405 compensate for slight electrical differences between the pulse transformer output windings to prevent the appearance of high pulse voltages across the filament.

The d-c plate current of the magnetron flows through the cathode circuit of *V*406, pins 3 and 5 of the pulse transformer, and the 220-ohm resistance *R*411 to ground. Coil *L*404 (1 millihenry) blocks peak pulses but passes average current through the magnetron current meter to ground. Capacitor *C*404 offers a low impedance to magnetron pulses and integrates them for metering. Negligible current is generated within the magnetron during quiescence, but the average or d-c current summed over a period of pulsed operation is of high magnitude. Longer pulse periods at lower PRF tend to balance shorter pulse periods at higher PRF, resulting in a nearly uniform duty cycle regardless of the radar function being performed. Note that while 115 volts is applied to the magnetron filament transformer during standby it is reduced to 65 volts when the high voltage is applied to the modulator.

VELOCITY-MODULATED TUBES AS LOCAL OSCILLATORS

In radar, most receivers use 30 or 60 mc. IF frequencies. In receiver operation, a highly important factor is the stability of the frequency of the local oscillator which generates the frequency that differs from the transmitted frequency by the IF. For example, in a radar receiver which receives a frequency of 3,000 mc., a frequency shift in the local oscillator as much as 0.1 percent would be a 3-mc. frequency shift. This is equal to the bandwidth in most receivers and would cause a considerable loss of amplification in the receiver.

Still another consideration in radar is a receiver which uses a crystal diode mixer. In this receiver the power required of the local oscillator is small, being only 20 to 50 milliwatts in the 3,000-mc. region. Due to the very loose coupling, only about one milliwatt actually reaches the crystal.

A third requirement of a local oscillator is that it must be tunable over a range of several megacycles. This is to compensate for changes in the transmitted frequencies and in its own frequency. In cases where the RF units are located in a remote position, it is desirable to mount the converter section of the receiver remotely. This insures that the IF frequency rather than the transmitted is cabled back to the receiver, and makes it necessary that the local oscillator be tuned from a position some distance away, preferably by varying the voltage applied to it.

Because the reflex-velocity-modulated tube or klystron meets these three requirements better than any other type of tube now in production, it is used almost exclusively for local oscillators in microwave radar receivers. The klystron is also frequently used in microwave signal generators or other applications where a source of microwave RF energy of low power is required. The following deals with its operation. In it the cavity resonators are treated as parallel resonant circuits, as was explained in the theory of the resonant cavity in chapter 5.

Theory of Operation

Originally klystrons contained two cavity resonators. When a signal was applied to the first resonator, the second resonator would also contain the same signal in an amplified form. Thus, by providing the tube with the proper external coupling a part of the output resonator's power could be returned to the input. This feedback caused the tube to function as an oscillator. BUNCHING (to be explained later) was produced by the first resonator and the output was taken from the second resonator. In the following discussion of the reflex klystron notice that the bunching is produced by the same cavity from which the output is taken.

In the circuit of the reflex klystron (fig. 6–37), note the arrangement of the electrodes and the voltages involved for operation. Electrons are emitted by an indirectly heated cathode. These electrons are attracted by the cavity grids (buncher grids) which are more positive than the cathode by the voltage E_a. The control grid is located between the cathode and the cavity grids and is for the purpose of controlling electron flow. It also has a positive potential which is usually about two or three hundred volts. The electrons emitted from the cathode travel toward the cavity grids at a velocity determined mainly by E_a. Most of the electrons pass through the control grid, the cavity grids, and continue on toward the repeller plate. After passing the cavity grids they approach a region where the electrical field opposes their motion since the repeller plate is negative with respect to the cathode by the voltage E_r. This voltage is variable and equals about minus 100 volts. This, in turn, makes the voltage from repeller to cavity grids three or four hundred volts and slows down the electrons causing them to come to a stop, after which they reverse direction and pass back through the grids.

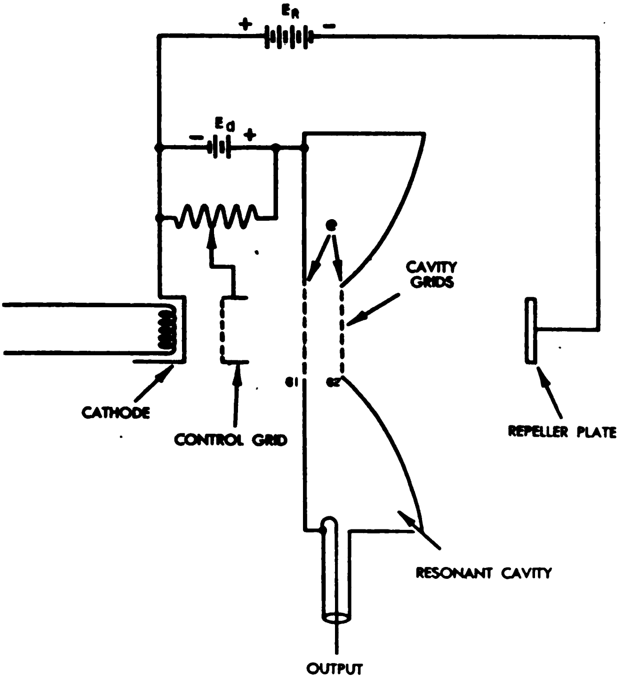


Figure 6–37.—Circuit of a reflex klystron oscillator.

Most of these returning electrons are collected by the cavity grid nearest the cathode, the accelerating grid, or the shell. Those that are not collected continue on toward the cathode until they are reversed in direction by the accelerating field and sent back through the cavity a third time. These electrons that are not collected cause improper tube operation; this undesirable feature is partially overcome through tube design.

It is important to consider the cause of oscillations in resonant cavities. In most oscillators, oscillations start from some irregularity in current flow, such as a transient that results from voltage being suddenly applied to the tube or from shot effect. With this in mind, assume that the oscillations in the resonant cavities are already taking place. From this assumption examine the source of the energy needed to sustain these oscillations.

With the tank circuit oscillating, a high-frequency voltage, e, appears between the two cavity grids. This causes the electric field between these grids to reverse twice during each complete cycle of operation. In this condition, as the electrons approach these grids, the electron stream is uniform. The time that is required for the electrons to pass through the small distance between the grids is small compared to the period of oscillations. Electrons which enter the space between the grids when e is zero will encounter no a-c electrical field and they will pass on through at the same velocity. The electrons which enter the space when e makes grid G_1 positive with respect to grid G_2 will encounter a field which tends to accelerate them. The amount of acceleration is a function of e. Electrons entering the space when e is reversed in polarity are decelerated. The change in velocity due to acceleration and deceleration is small in comparison with the original velocity. Those electrons which are accelerated most will travel farther toward the repeller plate before being turned back, while those that are decelerated most will be turned back sooner. Thus it is conceivable that with the proper magnitudes of e, E_a and E_R, electrons returning to the cavity grids will arrive in bunches.

Figure 6–38 shows the position of electrons in the tube at various times during their transit. The zero distance position, taken as the baseline in the graph, is midway between the cavity grids. Electron A, which arrives when G_2 is positive, is accelerated and travels farther before being turned back; electron B is unaffected; electron C is decelerated and turns back after a shorter excursion. Hence, in the diagram these electrons and the ones passing through at intermediate times are shown as arriving back at the grids at the same instant of time. This is the ideal situation, but it is not difficult to see that electrons will return to these grids in a stream which

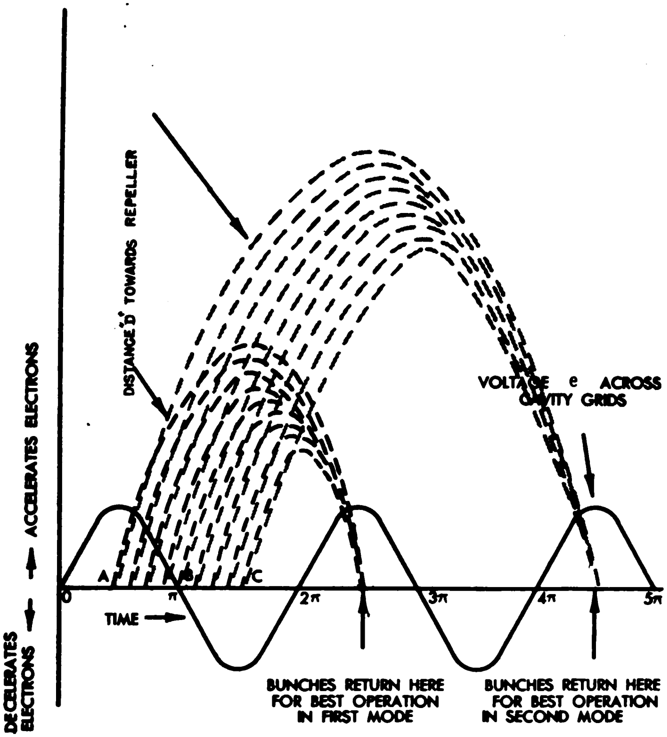

DISTANCE "D" TOWARDS REPELLER
VOLTAGE e ACROSS CAVITY GRIDS
DECELERATES ELECTRONS ← → ACCELERATES ELECTRONS
A
B
C
0
π
2π
3π
4π
5π
TIME →
BUNCHES RETURN HERE FOR BEST OPERATION IN FIRST MODE
BUNCHES RETURN HERE FOR BEST OPERATION IN SECOND MODE

Figure 6–38.—Bunching action of a reflex klystron.

varies in intensity at the frequency of the oscillations. Due to this fact this tube is called a REFLEX-VELOCITY-MODULATED tube because the electrons reverse direction and travel through the interelectrode space twice.

On the return trip the electric fields set up by the voltage *e* again act upon the electrons. Since they are now traveling in the opposite direction, they will be decelerated if they return when G_2 is positive and accelerated if they arrive when G_2 is negative. When an electron is accelerated by an electric field, its kinetic energy increases; this additional energy is taken from the electric field. On the other hand, an electron which is decelerated gives up energy to the electric field. If the bunches of electrons can be made to arrive back when G_2 is positive, they will give up energy to the alternating field. For maximum transfer of energy the bunches must arrive when G_2 is maximum positive. The question arises as to where this energy originates. Remember that if the electron stream from the cathode is uniform, some electrons are accelerated and some are decelerated on the outbound trip by the electric field of *e*. On the average, as many electrons absorb energy from the field as give up energy to it. Hence, very little net energy is taken from the oscillating circuit during the bunching process. The average kinetic energy of the electron is that imparted to it by the d-c voltage E_a. Thus, some of the energy taken from the d-c electric field is transferred to the a-c field to sustain the oscillations. The power output is taken from the klystron with an inductance loop located inside the resonant cavity.

Modes of Operation

It is not necessary that the bunches of electrons return to the grids on the first positive swing of G_2. Figure 6–38 indicates that if the bunches arrive on the second positive swing, the net result is still the same. You can see that the time in transit for the average electron *B* is ¾ cycle, 1¾ cycles, 2¾ cycles, and so forth. In actual practice there are three or four "modes" in which it is possible for the reflex-velocity-modulated tube (klystron) to oscillate. The dark lines in the illustration indicate paths of electrons operating in the first mode while the light lines indicate paths of electrons operating in the second mode. A third mode is possible when the average transit time is 2¾ cycles, and so forth.

The following shows how the transit time is controlled to produce oscillations in the different modes. Since the original velocity of an electron depends on the d-c voltage E_a, and since the distance that the electron travels before turning back and the speed with which it returns depend upon the difference between E_a and E_r, it is possible to adjust the two voltages E_a and E_r for any of the modes. The voltage E_a is usually fixed in magnitude since varying it produces greater initial velocity. This in turn would cause a farther excursion and a greater return velocity. E_r is made variable since it is not feasible to make E_a variable. For operation in the first mode, the round trip must be completed in the shortest time. This is accomplished by making the repeller plate highly negative. For longer time in the interelectrode space, the repeller is made less negative.

In figure 6–39, which shows power output and frequency of oscillations as functions of the repeller voltage for three modes of operation, notice that the frequency at the point of maximum output is the same for all three modes and is the resonant frequency of the cavity. In addition, note that the power output for the various modes at the resonant frequency are not the same and that the output is least in the highest mode. This can be explained by examining the factors which limit the amplitude of oscillations and which, in turn, limit the power output.

Power and amplitude limitations are due to overbunching as well as the usual losses in the oscillatory circuit. Overbunching occurs in the following way. As oscillations build up and e becomes greater, the amount of acceleration and deceleration increases. This causes bunching to occur within a shorter period of time, that is, in a time before the electrons reach the grids on the return trip. This tends to reduce the magnitude of oscillations. In the higher modes of oscillations where the bunches are formed more slowly, the electrons are more susceptible to overbunching. The magnitude of *e* which results in overbunching is therefore lower and oscillations are limited by this action to a lower amplitude than in the lower modes of operation.

As shown in figure 6–39, the frequency of oscillations in a reflex klystron is variable to a limited degree in any of the modes of operation by means of varying the repeller voltage. When the repeller voltage is varied, it causes a bunch to return a little sooner or a little later than normal. Off resonance, the amplitude of oscillations decreases by an amount which depends on the *Q* of the cavity. In this tube the tuning range is small in comparison to the frequency of oscillations and varies somewhat from one mode to another, being greatest in the highest mode. This can be explained from the fact that in the highest mode, bunching and debunching take place at a slower rate and that greater variation from the ideal time of return is possible without debunching, causing the amplitude of oscillations to drop below this usable output level. Another way to look at it is that in the highest mode the interval between leaving the grids and returning is greater, and the change in period represented by a given change in frequency is a smaller portion of the interval. To illustrate, in the third mode the interval before return must be about 2¾ cycles. A small change in the period of *e* would therefore be only ¾ divided by 2¾ or 3/11 as great a portion of the interval as it would if operation were in the first mode where the ideal time interval is ¾ cycle.

The band of frequencies which can be obtained by varying the repeller voltage lies between the half-power points shown in figure 6–39. This range of frequencies is known as the ELECTRICAL BANDWIDTH. The output

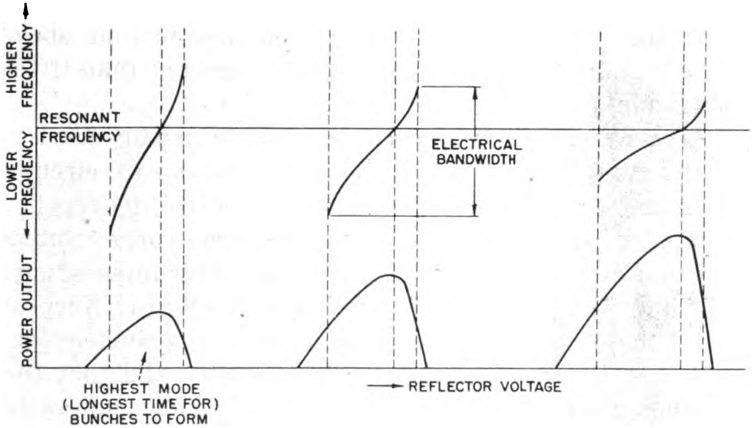


Figure 6–39.—Power output and frequency versus reflector voltage in different voltage modes for a reflex klystron.

curves of the bandwidth are unsymmetrical about the maximum output points. This results from the fact that if E_r is increased, not only does the bunching voltage e decrease and cause bunches to form at a later time, but the repeller voltage causes a quicker return and the effects of the two actions add to cause poor bunching at the time the electrons return, resulting in a rapid drop in output on the high side of the hump. At lower voltages, however, even though the bunching voltage e decreases and causes slower bunching, the decreased repeller voltage causes a later return to the grids so that the two effects are counteracting and a greater change in repeller voltage is possible before the output drops below the usable level.

The choice of the point and mode of operation is a compromise of several factors. To begin with, there are three or four modes which have the necessary power output. On the whole, it would seem then that the correct choice would be the highest mode, or it gives the largest tuning range. The highest mode, however, is too sensitive to a change in voltage to be well regulated. A change of one volt may cause a change of 0.5 mc. in the 3,000-mc. oscillator. Since the humps are unsymmetrical, the point of operation is usually chosen a little below the point of maximum output. This makes it possible to tune above the operating frequency by a greater amount than if the maximum point were used.

In a radar set, the reflex klystron is usually used in conjunction with an automatic-frequency-control circuit; a circuit which frequently is used to control the repeller voltage in such a way as to maintain the correct intermediate frequency. The details of the discriminator circuit and its operation are covered in chapter 4. Keep in mind that the frequency of oscillations is primarily determined by the dimensions of the cavity and that the repeller voltage is effective in making small changes in the frequency. Hence, in most reflex klystrons there is a coarse frequency adjustment which varies the cavity size in some way and the repeller voltage is the fine frequency adjustment. The dimensions of the cavity are changed (fig. 6–40) by a mechanical adjustment.

KLYSTRON THERMAL TUNING.—It has been found advisable in the later types of equipment to operate the klystron's local oscillator at a point of maximum output (klystron tuned for maximum crystal current). This is necessary because the noise output of the local oscillator is at a minimum at this point. Maximum output is obtained when the klystron's cavity is resonant to the output frequency. In order to accomplish this in a practical installation it is necessary that a tuning mechanism be used that will permit remote control of the cavity dimensions of the klystron.

One method of accomplishing this is through the use of a remotely controlled cavity tuning motor. A newer method of remotely tuning the klystron cavity has been incorporated in the design of some of the newer tubes. In this type tube the tuning mechanism is a part of the klystron. In one tube of this type the upper section contains a triode. (See fig. 6–40.) The plate of this triode changes in length in accordance with the tempera-

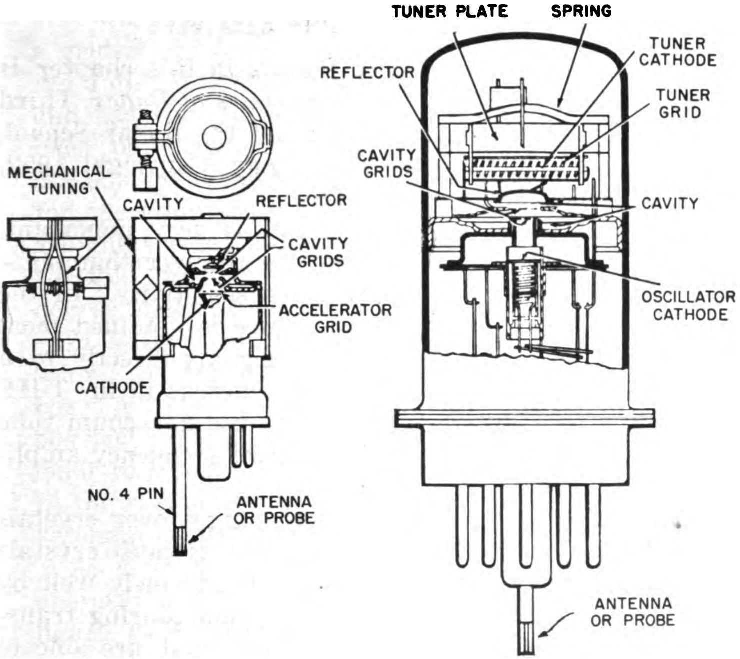


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Figure 6–40.—Construction of typical reflex klystrons.

ture of the plate. This temperature varies in accordance with plate current.

The variations of the triode's plate length are impressed through mechanical linkage on the cavity of the klystron. Thus, by controlling the grid voltage of the triode, and in turn its plate current, it is possible to vary the dimensions of the cavity in the klystron. This method of tuning is known as THERMAL TUNING.

In some tubes the triode is replaced by a diode. This diode has a control tube connected in series with it and a B supply. By varying the bias on the control tube it is possible to control the diode's plate current and this in turn varies the dimensions of the klystron's cavity.

INPUT CIRCUITS OF UHF RECEIVERS

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A careful choice of input circuit is very important, because it is in this portion of the receiver that noise has the greatest effect on receiver sensitivity. In the microwave region, a low noise figure is obtained most readily by applying the received signals directly to a mixer circuit employing a crystal, whereas in the UHF range it is usually advantageous to use a vacuum tube preceded by one or two stages of radiofrequency amplification.

Vacuum tubes have several advantages over crystals in the input circuits of a radar receiver. Because crystals are delicate, they must be protected extremely well by the TR device in order to avoid burnout during transmission. Vacuum tubes, on the other hand, are able to withstand relatively large amounts of transmitter power. Mixer circuits employing triodes also have a somewhat higher conversion gain than crystal mixers, although they generate more noise. The effect of the increased noise may be offset at ultrahigh frequencies by preceding the triode mixer by one or more stages of radiofrequency amplification—the noise figure of a triode is appreciably less when used as an amplifier than when used as a mixer.

Unfortunately, the tubes that are available for operation in the microwave region generate an excessive amount of noise and are therefore not suitable for the input circuits of microwave receivers. Consequently, use of vacuum tube input circuits is restricted to receivers that operate in the UHF range. First, vacuum tube input circuits are discussed; and second, several forms of crystal-mixer input circuits are discussed.

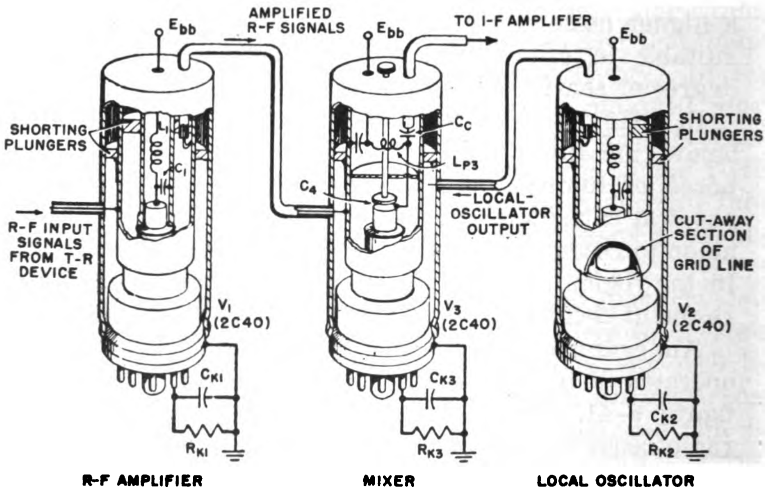
The input circuits of UHF receivers usually employ disk-seal triodes and are frequently connected to form a grid-return type of circuits. An RF amplifier employing a disk-seal tube in a grid-return circuit is very stable and has very little tendency to oscillate, the only feedback path being through the extremely low plate-to-cathode capacitance of the tube. Pentodes are available for operation in the UHF range, but they generally have a higher noise figure than triodes and are therefore less suitable in radar or other applications when a high degree of sensitivity is required.

In VHF receivers that operate at 200 megacycles or below, the tuned circuits of the input stages are composed of lumped inductances and capacitances. Often physical capacitors are unnecessary, the inductors being made to resonate with the tube and wiring capacitances. In the region of 1,000 megacycles the lumped-parameter resonant elements are replaced by cavity resonators.

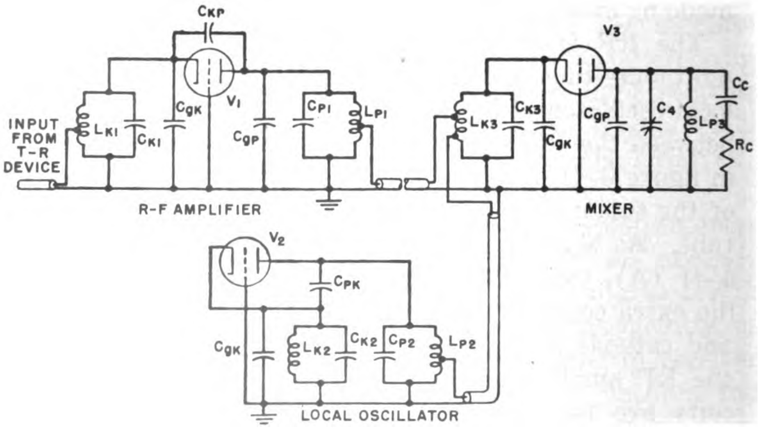
An example of an input circuit of a receiver that operates in the vicinity of 1,000 megacycles is given in figure 6–41. The circuit consists of a single stage of radiofrequency amplification, a local oscillator, and a mixer. Lighthouse tubes and coaxial-line resonators are employed in each unit, and connections between units are made by means of short sections of coaxial cable.

The RF amplifier is similar in construction to the lighthouse-tube oscillator shown in figure 6–5 and the local oscillator in figure 6–41. The only major difference between the RF amplifier and the local oscillator shown in figure 6–41 is that in the amplifier the middle cylinder of the tuner is not cut away at the grid terminal of the tube. As shown in the local-oscillator section of figure 6–41 (A), the cutaway section of the grid line provides the extra coupling needed in the oscillator between plate and cathode lines. Since there is no cutaway section in the RF amplifier's grid line, the plate and cathode circuits are isolated except for the small interelectrode capacitance between the plate and cathode of the tube.

The plate supply voltage is conducted to the plate terminal of the amplifier tube V_1 by means of an insulated wire which passes through the innermost coaxial cylinder. This cylinder, which serves as one conductor of the plate line, is coupled to the plate terminal through capacitor C_1 in order that the tuner may be operated at ground potential. Inductor L_1 serves as a choke to pre-



(A) CIRCUIT LAYOUT



(B) RF EQUIVALENT CIRCUIT

Figure 6–41.—Example of an input circuit of a UHF receiver.

vent RF signals from entering the power supply. The middle cylinder of the tuner connects directly to the grid terminal, and the outer cylinder fits over the shell of the tube, which is grounded. All cylinders are connected together by the shorting plungers in the plate and cathode lines. A bypass capacitor within the tube provides an RF connection between shell and cathode, and cathode bias voltage is developed across C_{k1} and R_{k1} in parallel between the cathode pin and ground. Input signals are coupled into the cathode-to-grid line by means of a probe, and a coupling loop is used to take the amplified RF signals from the plate-to-grid line.

An RF equivalent circuit of the amplifier, in which the plate and cathode lines are replaced by parallel L–C circuits, is shown in the upper left portion of figure 6–41 (B). Observe that the equivalent circuit is drawn so as to emphasize the grid-return aspect of the amplifier. The plate line, represented by L_{p1} and C_{p1} in the figure, is adjusted to resonate with the plate-to-grid capacitance C_{gp} of the tube at the frequency of the incoming signal, and the cathode line in parallel with C_{gk} is adjusted very nearly to the frequency of the incoming signals. Actually, the cathode line is made slightly inductive in order that any feedback voltage supplied through the small plate-to-cathode capacitance C_{pk} (approximately 0.05 $\mu\mu$f) will have a phase opposite to that required to start oscillations. Thus the cathode-line length is somewhat less than a quarter wave at the signal frequency.

Because the output coupling loop of the amplifier is located near the shorting plunger in the plate line, the output coaxial cable is shown as a tapped connection to L_{p1} in figure 6–41 (B). The plate line thus acts as a resonant transformer to change the low input resistance of the cable (approximately 50 ohms, since the cable is usually matched at the far end) to a higher resistance across the entire resonant circuit. The value of load resistance across the resonant circuit is governed primarily by the bandwidth required for the amplifier.

Since no provision is made for tuning the amplifier during its normal operation, the bandwidth must be wide enough to pass received signals despite slight changes in the transmitted frequency.

The tuned cathode circuit is loaded by the input resistance of the amplifier in parallel with a resistance representing grid-loading effects and tuned-circuit losses. The net resistance is matched approximately to the characteristic resistance of the input line by tapping the input cable to a point near the shorted end of the cathode line. The tap location is chosen for optimum signal-to-noise ratio and results in a resistance termination for the input line which is slightly more than its characteristic resistance.

The local oscillator, shown at the right in figure 6–41 (A), is similar in construction to the RF amplifier except that the grid cylinder is partly cut away near the local-oscillator tube. Thus C_{pk} in the oscillator equivalent circuit (fig. 6–41 (B)) includes both the plate-to-cathode capacitance of the tube and the external coupling capacitance resulting from the opening in the grid line. All cylinders of the oscillator tuner, like those of the RF amplifier, operate at ground voltage, and the cathode is maintained at a positive potential with respect to ground because of the bias voltage developed across C_{k2} in parallel with R_{k2}.

The local oscillator is tuned in accordance with the principles discussed for the lighthouse-tube oscillator in the early part of this chapter. The plate-line length is the primary control for frequency and, in fact, constitutes the main tuning control for the receiver. Tuning is accomplished by means of a knob or dial connected to the plate-line shorting plunger through a mechanism not shown in the figure. The cathode-line length governs the magnitude and phase of the voltage fed back to the cathode circuit from the plate circuit and must be adjusted so that a capacitive reactance appears between grid and cathode at the oscillation frequency. At frequencies in the neighborhood of 1,000 megacycles per second, it is mechanically feasible to choose line lengths so that each line employs its lowest mode of resonance. Thus the cathode line is approximately one-quarter wave at the oscillator frequency, and the plate line is less than a quarter wave.

Power coupled from the plate line of the local oscillator is applied, together with echo signals from the RF amplifier, to the cathode line of the mixer tube V_3. A component of current at a frequency equal to the difference between the local-oscillator and echo-signal frequencies is produced in the plate circuit of the mixer, and the plate circuit is tuned to this difference frequency.

The mixer tuner differs from the local-oscillator and RF amplifier tuners in that it employs a coaxial-line cavity only between cathode and grid, lumped parameters being used in the plate circuit. The tuned-plate circuit consists of inductor L_{p3} in parallel with interelectrode capacitance C_{gp}, the series combination of C_c and the input resistance R_c of the output line, and the variable capacitance C_4 formed by a movable metallic disk near the plate terminal of the tube. The capacitance of C_4 is varied by changing the disk position by means of a rod which passes through the end of the tuner.

The various components comprising the tuned-plate load of the mixer are illustrated in figure 6–41 (B). Because input signals are applied between grid and cathode of the mixer and the intermediate-frequency signals appear between grid and plate, the equivalent circuit is represented as a grid-return circuit. From the equivalent circuit it may be seen that the variable capacitor C_4 serves to tune the remaining plate-circuit elements to resonance at the intermediate frequency. Placing capacitor C_c in series with the output cable isolates the cable from the power supply voltage and causes the low input impedance R_c of the cable to appear as a higher resistance across the entire circuit.

The receiver input circuit of figure 6–41 (A) requires that the interconnecting cable between RF amplifier and mixer be short. Otherwise appreciable attenuation of received signals will occur in the cable and cause noise generated in the mixer to become important. If cable loss is excessive, it may become necessary to employ a second stage of RF amplification in order to maintain the signal level well above the noise level of the mixer.

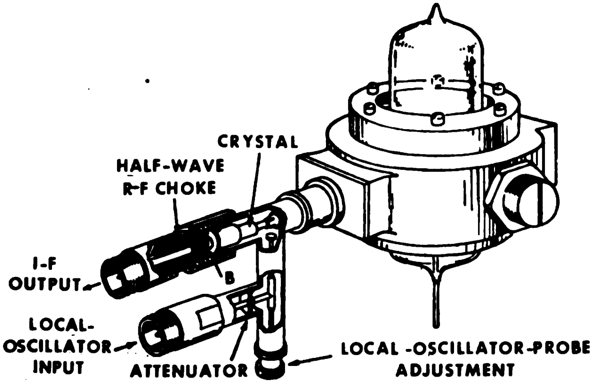
CRYSTAL MIXERS

Single-Ended Crystal Mixers

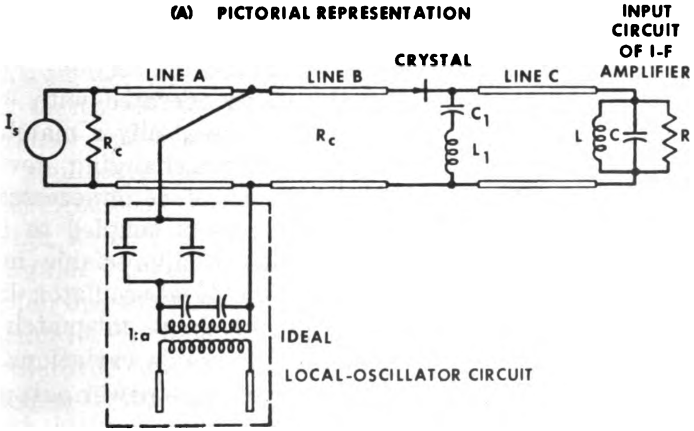
In microwave receivers, amplification of received signals at the carrier-frequency level is in general undesirable because of the rather large amount of noise generated by available microwave-amplifier tubes. Echo signals of a microwave radar system, after passing through the TR device, are therefore applied directly to the crystal mixer. A voltage from a local oscillator, usually a klystron, is applied to the mixer simultaneously with the echo signals, and signals at the intermediate frequency are derived at the mixer output terminals.

The important features of a 10-cm. mixer employing coaxial-line sections are given in figure 6–42 (A). Echo signals are coupled into the transmission-line section between TR cavity and crystal by means of a coupling loop which penetrates a short distance into the cavity. The crystal fits snugly into the inner conductor of the line and is held concentric with the outer conductor by means of a dielectric bead. The local-oscillator signal is introduced into the mixer assembly by means of a coupling probe located approximately a quarter wave (at the local-oscillator frequency) from the coupling loop in the TR cavity. A high impedance in the direction of the cavity is thus presented to the local-oscillator signal at the probe. The strength of the local-oscillator signal at the crystal may be changed by varying the distance the probe extends into the main line. Power from the local oscillator is conducted to the proble over a section of transmission line.

Connection between the crystal and inner conductor



(A) PICTORIAL REPRESENTATION



(B) AN EQUIVALENT CIRCUIT

Figure 6–42.—A 10-cm. mixer.

of the line section carrying IF signals is provided by a spring contact. The half-wave RF choke appearing on the IF side of the crystal provides an electrical short circuit for radiofrequency signals at point *B*. The d-c return path required for crystal current is provided at the RF end of the mixer assembly by the TR coupling loop and at the IF end by the shunt-inductance element located in the grid circuit of the input IF amplifier stage.

A two-wire equivalent circuit of the mixer is given in figure 6–42 (B). If small losses in the RF line and TR cavity are ignored, the portion of the RF transmission system between antenna and output terminals of the TR cavity may be represented by a signal-current source I_s in shunt with the characteristic resistance R_c of the transmission line. Because the local oscillator is loosely coupled to the main line, negligible admittance appears in shunt with the lines *A* and *B* in figure 6–42 (B). Elements L_1 and C_1 represent the RF choke at the IF end of the crystal and are series resonant at the echo-signal frequency. Hence line *B* is terminated by the crystal at the echo-signal frequency. For maximum receiver sensitivity it is important that essentially all the echo power available at the TR device be transferred to the crystal. The crystal should therefore present a matched-load termination to the main-line section. Fortunately, 10-cm. crystal units, when operated with 5/16-inch 50-ohm coaxial lines, provide essentially a matched RF load for the line. An impedance-changing device between crystal and TR cavity is therefore unnecessary.

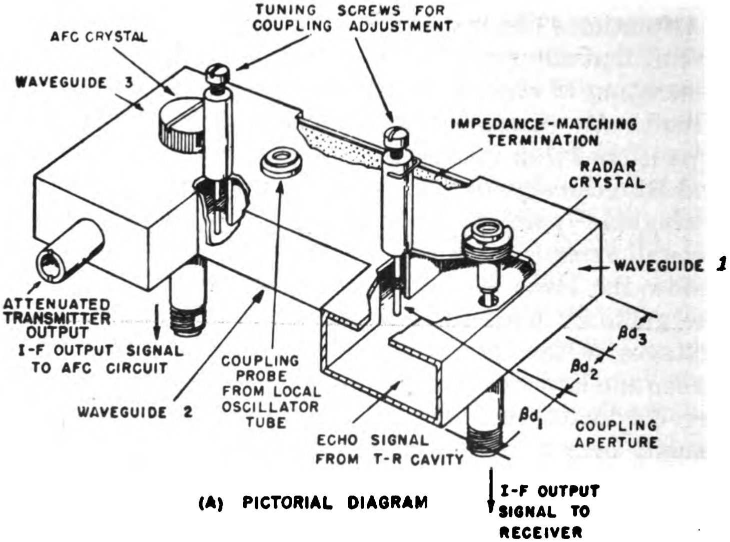
Because the local oscillator is loosely coupled to the main line containing the crystal, an appreciable mismatch of impedances exists on the local-oscillator line at the coupling probe. A large impedance mismatch is in general undesirable because of the wide variations in local-oscillator load impedance (and hence power output) that result when the oscillator frequency is varied. A good impedance match is particularly important if the line between local oscillator and probe is long. To insure a relatively constant load impedance for the oscillator, a resistance-disk type of attenuator is oftentimes inserted in the local-oscillator line, as shown in figure 6–42 (A), in order to terminate the line section between oscillator and disk in its characteristic resistance. An alternative to the resistance disk is a piece of lossy cable which may be inserted between local oscillator and mixer assembly. (NOTE: Lossy cable uses resistance wire for the center conductor. An example of lossy cable is RG–21/U.)

Oftentimes mixer asemblies include, in addition to the circuit that supplies IF echo signals, a separate mixer for generating IF signals for the automatic-frequency-control circuit. An example of a 3-cm. mixer assembly of this type is shown in figure 6–43 (A). Local-oscillator power and RF echo signals from the TR cavity are coupled into waveguide 1, which serves as the echo-signal channel. A crystal extends between top and bottom walls of the guide, the lower end of the crystal being isolated from the guide by means of a dielectric washer. Echo-signal voltages in the intermediate-frequency band appear between the lower end of the crystal and the waveguide and are conducted to the input circuit of the IF amplifier, usually over a coaxial cable.

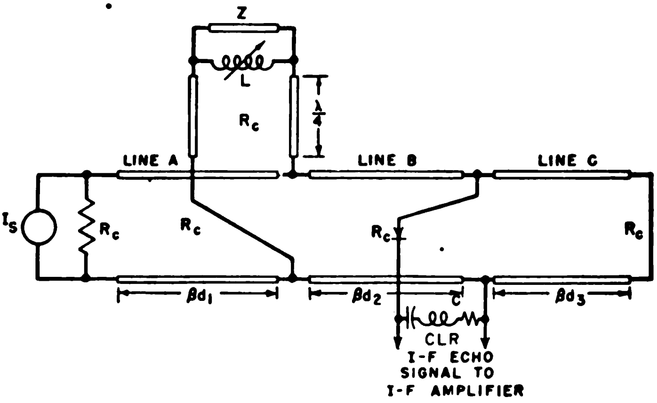
Waveguide 3 is similar in construction to waveguide 1 and serves as the AFC channel. Power from the local oscillator and a small amount of the transmitter power are applied to the AFC crystal, and IF signals for the AFC circuit are derived at the lower end of the crystal mount, as before. Attenuation between the main guide and guide 3 is provided by means of a circular-waveguide section operating below cutoff. Sometimes a disk type of attenuator is included in the circular section to reduce further the amount of transmitter power reaching the mixer assembly.

The local-oscillator tube is mounted above waveguide 2 in a manner that permits the local-oscillator output probe to extend a short distance into guide 2. The coupling probe may be displaced slightly from the center of the transverse axis of the guide in order to obtain a load impedance for the oscillator which yields a reasonably uniform power output over the tuning range.

Local-oscillator power is coupled from guide 2 into guides 1 and 3 through coupling apertures located in the side walls of the guides. A matching post and tuning screw extending from top and bottom walls of the guides at the aperture locations serve to regulate the amount of local-oscillator power coupled to the mixer circuits.


TUNING SCREWS FOR COUPLING ADJUSTMENT
AFC CRYSTAL
WAVEGUIDE 3
IMPEDANCE-MATCHING TERMINATION
RADAR CRYSTAL
WAVEGUIDE 1
ATTENUATED TRANSMITTER OUTPUT
I-F OUTPUT SIGNAL TO AFC CIRCUIT
COUPLING PROBE FROM LOCAL OSCILLATOR TUBE
WAVEGUIDE 2
ECHO SIGNAL FROM T-R CAVITY
βd3
βd2
βd1
COUPLING APERTURE
I-F OUTPUT SIGNAL TO RECEIVER

(A) PICTORIAL DIAGRAM



(B) AN R-F EQUIVALENT CIRCUIT OF ECHO-SIGNAL CHANNEL

Figure 6–43.—A 3-cm. mixer.

For optimum performance of the radar system, the echo-signal channel of the mixer should provide a matched-load impedance to the waveguide transmission system. The elements that determine the mixer input impedance are shown in the equivalent circuit of figure 6–43 (B). The equivalent circuit is derived by dividing guide

1 into three sections of lengths βd_1, βd_2, and βd_3 and representing the sections by equivalent lengths of transmission lines (lines *A*, *B*, and *C* in fig. 6–43 (B)). The signal-current generator I_s in shunt with R_c in the equivalent circuit accounts for the portion of the RF system between antenna and mixer. The coupling aperture between local-oscillator and mixer guides (guides 1 and 2) is represented by a variable inductance *L*, and in shunt with *L* is an impedance *Z* which accounts for the impedance seen in the direction of the local-oscillator guide. The parallel combination of *L* and *Z* terminates a quarter-wave section of line which is in shunt with lines *A* and *B*. At the echo-signal frequency, the impedance of *L* and *Z* in parallel is maintained small enough in comparison with the characteristic resistance of the quarter-wave line to make the input impedance of the quarter-wave section extremely high and hence the amount of echo-signal power entering the local-oscillator guide extremely small.

The crystal is located at the junction of lines *B* and *C*, as shown in figure 6–43 (B), and across the crystal output terminals in a series *R–L–C* circuit. The counterpart of the *R–L–C* combination in figure 6–43 (A) is an RF choke in the IF output lead of the mixer. The choke prevents RF signals from entering the input circuit of the IF amplifier.

The length of shorted-end line *C* must be such that the input admittance of line *C*, in shunt with the RF admittance of the crystal, provides essentially a matched load for line *B*. Varying the length of line *C* changes the susceptance component of the total admittance at the crystal location. A length of approximately $\lambda/4$ is usually preferable in order to minimize changes in admittance with frequency.

Balanced Crystal Mixers

The noise introduced into the input circuit of a microwave receiver by the local oscillator may be reduced appreciably through use of a balanced mixer. An input circuit of this type employs two crystal mixers connected so that the local-oscillator voltages applied to each crystal are in phase and the echo-signal voltages at each crystal are 180 degrees apart in phase. As explained below, this method of circuit connection causes the IF components of local-oscillator noise to cancel in the IF amplifier input circuit and the IF components of the signal to add. Because of their ability to suppress local-oscillator noise and because of certain additional advantages to be discussed later, balanced mixers are oftentimes preferable to single-ended mixers. They are presently finding wide use in microwave radar systems.

The basic principles of a balanced mixer may be explained from the diagram of figure 6–44, which shows two crystals mounted a half wavelength apart (at the signal frequency) between conductors of a coaxial-line section. Each crystal is located a quarter wave from the nearer end of the line, and the local-oscillator signal is probe-coupled to the line at a point midway between the crystals. The output terminals of the crystals are connected to an IF transformer, the primary winding of which is balanced to ground. The signal-current generator I_s in shunt with resistance R_c accounts for signals entering the mixer from the TR device.

The local-oscillator coupling probe gives rise to two waves traveling in opposite directions in the coaxial line. Assume that at the local-oscillator frequency the TR device presents a short circuit to the local-oscillator wave traveling to the left in figure 6–44 and that the crystals have identical RF impedances equal to twice the characteristic resistance of the line. Then equal-amplitude, in-phase, local-oscillator voltages appear at the crystals. Because of the 180-degree spacing between crystals, the RF echo-signal voltages at the crystals are also of equal amplitude but are 180 degrees apart in phase. An in-phase relation between local-oscillator and echo-signal voltages produces an IF signal voltage of one phase and a 180-degree phase relation between local-oscillator and echo-signal voltages produces an IF signal voltage of

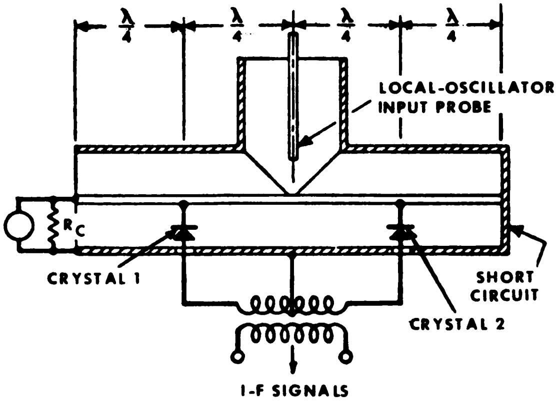
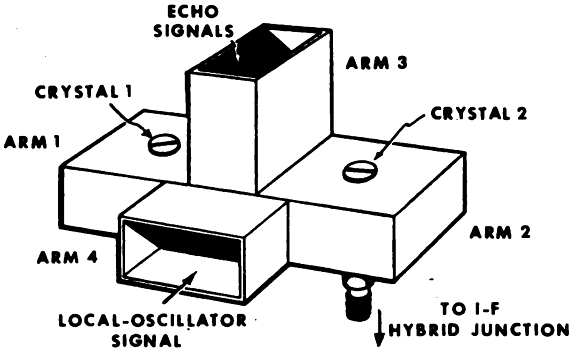


Figure 6—44.—Balanced-mixer circuit.

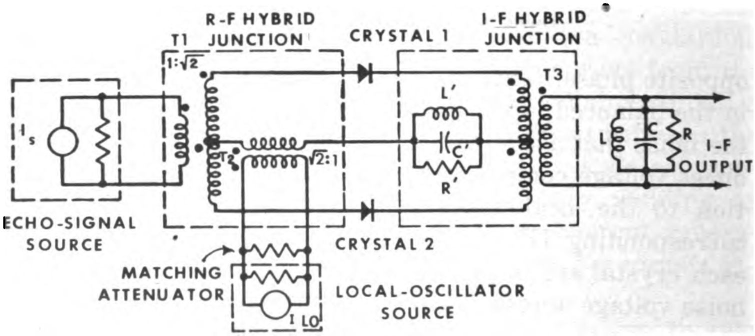
opposite phase. Thus the two IF echo-signal voltages add in the balanced input circuit of the IF amplifier. Because the noise that accompanies the local-oscillator signal produces voltage components that bear the same phase relation to the local-oscillator voltage at each crystal, the corresponding IF components of noise at the output of each crystal are in phase and hence produce no resultant noise voltage across the IF transformer.

In order to realize full benefit from use of a balanced-mixer circuit, the crystals should have similar RF impedances as assumed above, and in addition they should have similar conversion gains. Under these conditions of operation, only one-half the available RF power is applied to each crystal, but since the IF signal powers from the crystals add together, the total IF power is the same as that obtained from a single-ended mixer. Since IF components of local-oscillator noise are completely eliminated, maximum reduction in overall noise figure is obtained.

An unbalance in conversion gain yields an unbalance in the IF signal voltages obtained from each crystal and also an unbalance in the IF components of local-oscillator-noise voltages at the crystal output terminals. However, since the IF signal voltages are additive in the crystal output circuit and the local-oscillator-noise components



(A) PICTORIAL DIAGRAM



(B) EQUIVALENT CIRCUIT

Figure 6–45.—Balanced mixer employing hybrid junctions.

are subtractive, small unbalances in conversion gains do not seriously affect the overall receiver noise figure.

More important are the effects of a difference if the RF admittances of the crystals. An unbalance of this type causes both the echo-signal power and the local-oscillator power to divide unequally between the two crystals. As a result, the IF signal power obtained from the mixer is appreciably reduced, and in addition the IF components of local-oscillator noise are prevented from canceling completely in the IF input circuit. Crystals intended for use in a mixer constructed as shown in figure 6–44 must be carefully selected on the basis of equal

RF admittances in order to insure most satisfactory operation.

A balanced mixer that is superior in several respects to the coaxial-line mixer just described is shown in figure 6–45 (A). This mixer employes a waveguide hybrid junction of the type discussed in chapter 5; this junction is often called a MAGIC T. Crystals are located in one pair of entries (arms 3 and 4). Usually in a mixer of this type the input circuit of the IF amplifier is also arranged in the form of a hybrid junction. Details of connections of the IF amplifier input circuit, together with the ideal-transformer equivalent circuit of the waveguide junction are given in the right-hand block of figure 6–45 (B). The parallel circuit comprising *L*, *C*, and *R* is tuned to the intermediate frequency and the impedance of the parallel combination of L', C', and R' is made equal to the conjugate of the IF impedance appearing between the terminals to which L', C', and R' are connected.

Ideally the two crystals should be alike. Furthermore, all RF impedances should terminate the entries of the RF hybrid junction in matched loads, and all IF impedances should similarly terminate the entries of the IF hybrid junction. Then, because of the properties of a matched hybrid junction, local-oscillator signals at the crystals are equal in amplitude and in phase, and echo signals at the crystals are equal in amplitude and 180 degrees apart in phase.

The IF components of signals from the crystals are applied in push-pull to ideal transformer T_3 and appear across *L*, *C*, and *R* but not across L', C', and R'. The IF components of noise, on the other hand, are in effect applied between end terminals of T_3 in parallel and appear across L', C', and R' but not across the IF output terminals. Local-oscillator noise is thus suppressed completely. Because the RF hybrid junction is matched, the total available power from the echo-signal and local-oscillator arms divides equally between the two crystals. As in the coaxial-line mixer of figure 6–44, the IF power from each crystal is one-half the available echo-signal ~~power times the conversion gain G of the crystal.~~ Since IF power from each crystal is delivered at the IF output terminal, the total IF output is the product of the echo-signal power and G, as before.

Ordinarily, in single-ended mixers and in the coaxial-line balanced mixer (fig. 6–44), the local oscillator must be loosely coupled to the mixer in order to prevent an appreciable amount of echo-signal power from entering the local-oscillator circuit. As a result, the local oscillator must generate much more power than is actually needed for mixer operation, the excess power being dissipated in an impedance-changing device located in the local-oscillator circuit. Because in a hybrid-junction mixer the echo-signal and local-oscillator arms are decoupled, there is no tendency for echo signals to enter the local-oscillator circuit, and hence there is no need for loose coupling between local oscillator and mixer. A local oscillator with relatively low-power output can therefore be used, the only power necessary being that which is needed to drive the crystals. Actually most local oscillators are capable of generating more than the minimum power required, so that a dissipative attenuator can be inserted in the local oscillator in order to insure a matched termination for the arm. Such an attenuator is desirable, as will be seen later.

Decoupling between echo-signal and local-oscillator arms also means that there is no tendency for local-oscillator power to enter the echo-signal arm. Rejection of local oscillator signal by the echo-signal arm is particularly advantageous in systems employing a broadband TR device which would allow local-oscillator power to pass through and be radiated by the antenna.

Another advantage of the hybrid-junction mixer is that its operation is less sensitive to differences in the RF impedance of the crystals than is the balanced coaxial-line mixer. Because the two crystal arms are decoupled, any RF power reflected from one crystal cannot reach the other crystal. All the reflected power is directed into the echo-signal and local-oscillator arms and absorbed by the matched terminations. Thus the amount of RF power delivered to one crystal is independent of the RF impedance of the other.

If a mismatch should exist in either the echo-signal or local-oscillator arm, a reflection of an RF signal arriving at the mismatched termination will occur, and the signal reflected from the termination will enter the two crystal arms. Under these circumstances, the two crystal arms are not completely decoupled. At the echo-signal frequency, a mismatch in the echo-signal and local-oscillator arms is unlikely because in general the TR device provides a matched load for the echo-signal arm and, as indicated above, a matching attenuator can normally be inserted in the local-oscillator arm.

In general, if the crystals provide an impedance match at the echo-signal frequency, they do not yield a perfect match at the local-oscillator frequency. This mismatch is unimportant, since the reflected power will be absorbed in the echo-signal and local-oscillator arms. The attenuator in the latter arm serves to prevent the returned signal from reacting on the oscillator. Should the echo-signal arm not provide a matched termination for the returned local-oscillator signal, further reflection of the signal will occur and the balance of local-oscillator power at the crystals will be further upset. The effect of such an unbalance is to change the conversion gains of the crystals slightly, but usually the change is negligible even with moderate amounts of mismatch at the local-oscillator frequency.

Use of the hybrid-junction type of IF input circuit shown in figure 6–45 (B) insures suppression of local-oscillator noise despite unbalance in the IF impedances of the crystals. When all entries of the IF hybrid junction are terminated in a load that is the conjugate of the impedance the load faces, the two crystals, constituting one pair of entries, are decoupled and the two resonant circuits, constituting the other pair, are also decoupled. If the resonant-load entries remain matched, the amount of IF signal power coupled from one crystal to the output terminals is independent even if the IF impedance of the other crystal does not match the arms. Furthermore, the IF components of noise cancel in the output circuit independently of the IF crystal impedance as long as the crystals develop equal IF noise voltage. The IF components of noise appear only across the L', C', and R' arm. If crystals are tested for equal conversion gains while loaded by the conjugate of the IF impedance of a typical crystal, equal IF noise voltages can be expected when the crystals are employed in the circuit of figure 6–45.

Terminating the IF output entry in a matched load does not in general insure minimum noise figure. Mismatching the output entry in order to minimize the noise figure alters the load for the push-pull IF echo signals and thus reduces the echo-signal power delivered to the load but does not alter the load for the IF components of noise which are applied to T_3 in parallel. The local-oscillator noise-suppression property of the mixer is therefore retained.

QUIZ

1. The grid and plate tanks of the tuned-grid tuned-cathode oscillator are tuned
   a. above the oscillator frequency
   b. below the oscillator frequency
   c. to the oscillator frequency
   d. to resonance

2. To increase feedback in the tuned-grid tuned-cathode oscillator you would tune the cathode tank
   a. more capacitive
   b. less capacitive
   c. inductive
   d. resistive

3. In the lighthouse oscillator, the oscillator frequency is
   a. above the resonant frequency of the plate tank
   b. below the resonant frequency of the cathode tank
   c. same as the resonant frequency of the plate tank
   d. below the resonant frequency of the plate tank

4. The lead inductance which produces degeneration at VHF and UHF operation is
   a. plate-lead inductance
   b. grid-lead inductance
   c. cathode-lead inductance
   d. all of the above

5. The consumption of power due to the transit time of a tube, which occupies an appreciable portion of the input signal, is effectively the same as adding
   a. a resistance between grid and cathode
   b. a resistance between plate and cathode
   c. a resistance between plate and grid
   d. none of the above

6. The RF equivalent of the ultraudion oscillator is that of a/an
   a. Hartley oscillator
   b. Armstrong oscillator
   c. Colpitts oscillator
   d. Butler oscillator

7. The most common type of magnetron used today is the
   a. traveling-wave or resonant cavity type
   b. split-anode electron resonance type
   c. split-anode negative resistance type
   d. double split-anode type

8. The cavities of a traveling-wave type of magnetron are connected so that they appear to the output as
   a. series connected
   b. series, parallel connected
   c. separate outputs
   d. parallel connected

9. The method by which the magnetron sustains its oscillations is the
   a. electrons giving up energy to the d-c field
   b. electrons giving up energy to the RF field
   c. electrons giving up energy to the magnetic field
   d. stopping of the electrons by the tube's plate

10. Straps are used in a magnetron to
    a. remove power from the cavities
    b. increase frequency stability
    c. decrease the coupling between cavities
    d. excite additional modes of operation

11. One of the functions of a radar modulator is that it
    a. selects the timing sequence of the radar
    b. contains all the timing circuits of the radar
    c. determines the waveform of the output pulse of the magnetron
    d. supplies normally a positive pulse to the magnetron

12. The mixer stage of a radar receiver is a
    a. germanium crystal
    b. quartz crystal
    c. vacuum tube
    d. linear device

13. The hybrid junction mixers
    a. cancel local oscillator noise completely
    b. reinforce the IF
    c. provide better signal-to-noise ratio
    d. are more fragile than other types

14. Crystal current is adjusted to its optimum value by
    a. a resistive disk-type attenuator
    b. local-oscillator tuning
    c. local-oscillator coupling
    d. a lossy cable

15. For best operation of balanced mixers
    a. both crystals should be matched
    b. it makes no difference as to the match
    c. both crystal currents should be equal
    d. both crystals should have the same RF impedance

16. A klystron has
    a. one mode of operation
    b. two possible modes of operation
    c. three or more modes of operation
    d. only one mode of operation for each frequency

17. Klystrons can be tuned
    a. electronically with the buncher grids
    b. electronically with the repeller voltage
    c. by changing the distance from plate to cathode
    d. only at the factory

CHAPTER 7

UHF AND MICROWAVE ANTENNAS

The frequencies which are included in the UHF and microwave ranges extend from 225 megacycles to approximately 30,000 megacycles. These frequencies are divided into bands and are identified by the letters P, L, S, X, and K. The following table shows the approximate frequency range for each band and its letter designator:

P—225 to 390 mc.

L—390 to 1,650 mc.

S—1,650 to 5,200 mc.

X—5,200 to 10,900 mc.

K—10,900 to 33,000 mc.

The UHF range is included in the frequencies of the L and P bands; and the microwave range is included in the S, X, and K bands. The S and X bands are frequently referred to by wavelength instead of frequency. Since the S band is centered around the wavelength of 10 cm., it is referred to as the 10-cm. band, likewise the X band is referred to as the 3-cm. band. It is important to distinguish between UHF and microwave equipments since they possess many characteristics that differ because of their frequency of operation.

The fundamental principles of antennas for use at UHF and microwave ranges are the same as those for use at lower frequencies. These principles have been covered in the Navy Training Course, *Basic Electronics,* NavPers 10087, and will not be included in this chapter. It is advisable that you become thoroughly familiar with the material in the basic course for without such information you will most probably experience a great deal of difficulty in understanding this chapter.

The major purpose of a transmitting antenna is to convert the transmitter's RF power into radiated energy as efficiently as possible, and the major purpose of a receiving antenna is to absorb as much of this radiated energy as possible. In order to increase the effectiveness of the antenna system, several elements are often combined to form an ARRAY. The antenna array used will be determined by the requirements of the equipment and the particular purpose of the transmission. UHF directive arrays containing several parasitic elements may be built into a very small space. Likewise, directive arrays with parabolic and horn-type reflectors which would be too bulky for use at the lower frequencies become very practical at microwavelengths because of their small physical size.

ANTENNA ARRAYS

Driven Arrays

The driven array consists of two or more elements with all elements connected to the generator. They may be divided into three basic types—the broadside array, the end-fire array, and the collinear array.

BROADSIDE ARRAY.—When two half-wave elements are placed one-half wave apart, parallel to each other and excited in phase, most of the radiation takes place in a direction perpendicular to a plane through the elements. The arrangement and radiation pattern of a broadside array are shown in figure 7–1. In it, the radiation pattern is shown in solid lines. Increasing the number of elements makes the pattern more directional. This increased directivity is shown by the dotted lines.

END-FIRE ARRAY.—When two elements are spaced a certain fraction of a wavelength apart and are excited out of phase by the same fraction of a cycle, the radiation is directional in the plane of the array and per-

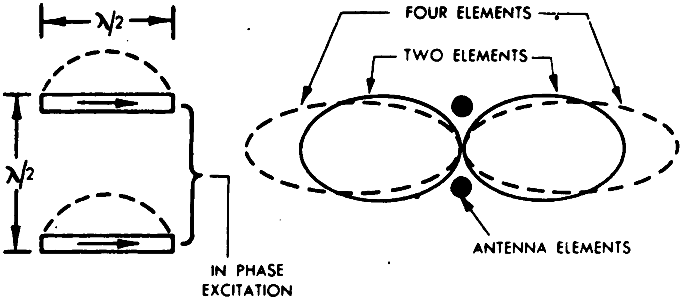


Figure 7–1.—Broadside array.

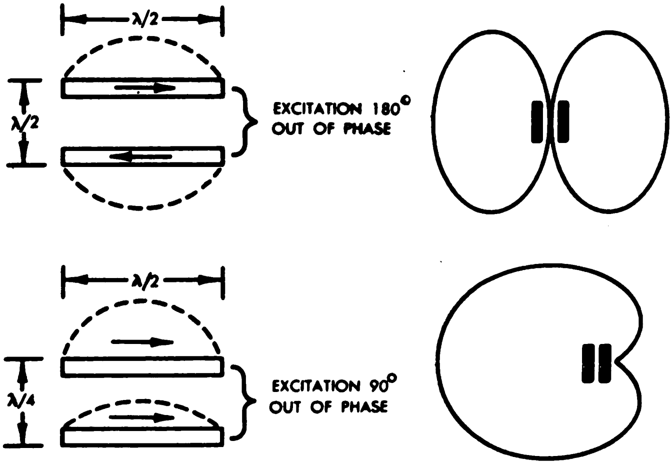


Figure 7–2.—End-fire array.

pendicular to the elements. It is directional from the end of the array that has the most lagging current. If the spacing in an end-fire array, for example, is a half wavelength, the two elements are excited a half cycle or 180 degrees out of phase. This causes the bidirectional pattern shown in figure 7–2. A unidirectional cardioid pattern can be obtained with a half wavelength spacing and a quarter cycle (90 degrees) phase difference in excitation.

COLLINEAR ARRAY.—A collinear array is formed when two half-wave elements are placed end to end and excited

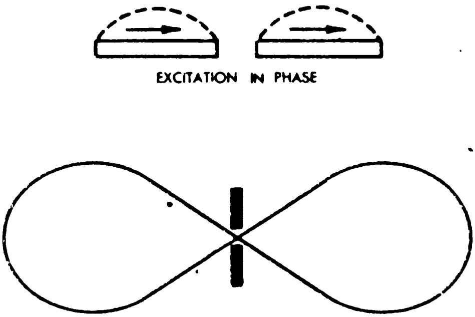


Figure 7–3.—Collinear array.

in phase. In a collinear array there is no directivity in the plane perpendicular to the antenna, but there is a sharp pattern in any plane containing the antenna. As with any of the other arrays, increasing the number of half-wave elements increases the directivity of the pattern. This array is shown in figure 7–3.

Parasitic Arrays

A parasitic array is an antenna system which consists of two or more elements in which only one of the elements is driven. The other element (or elements) is excited by induction and radiation fields which are produced by the driven element. With parasitic arrays it is possible to obtain highly directional patterns. The action in a parasitic array is analogous to the action in a transformer in which the primary induces a current in the secondary and the current in the secondary produces a magnetic field, which, in turn, induces current back into the primary. The phase relationship between elements in an array varies according to the spacing between elements. The elements are usually spaced an appreciable part of a wavelength apart.

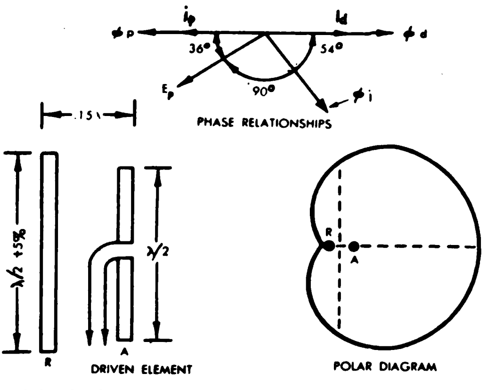
In a two-element parasitic array, shown in (A) of figure 7–4, the driven element (labeled A) is cut at the center for connecting a low impedance feed line. The length of the driven element is a half wavelength. This makes it self resonant. The parasitic element called the reflector is located 15 percent of a wavelength in space from the driven element. It is about 5 percent longer than the driven element.

Another diagram, (A) of the figure, shows the phase relationships in the parasitic array vectorially. Vector i_d, which represents the current in the driven element, is in phase with the H-field. The part of the H-field that cuts the parasitic element lags the field which leaves the driven element by 0.15 of a cycle. (This is the time lapse during the travel between elements.) The lag is equivalent to 54 degrees. The flux at the reflector is shown by vector Φ_i. It lags i_d by 54 degrees. The voltage induced by the field is 90 degrees out of phase with the field. This voltage is represented by the vector E_p. If the parasitic element were resonant, the current in it would be in phase with E_p, but the reflector is longer than the resonant half wavelength. A long antenna is inductive and the current in this element will lag the voltage by 36 degrees if it is approximately 5 percent longer. The radiated field will be in phase with this current. In summary, the field starting out from the parasitic element will be 180 degrees out of phase with the field leaving the driven element.

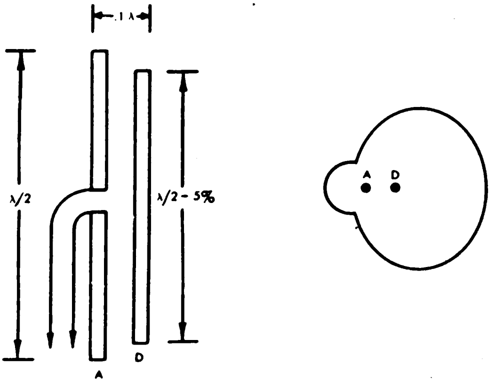
If the polar diagram for two elements spaced 0.15 wavelength apart and excited out of phase is plotted, the curve shown in (A) results. This shows that most of the radiation is on the side of the driven element which is away from the parasitic element, while very little occurs on the side of the reflector.

Another point is that the field passing from the reflector cuts the antenna (driven element) and induces a voltage in it. This voltage changes the input current. The input impedance, which is a function of this current, is about 50 ohms as compared to 73 ohms for the antenna alone.

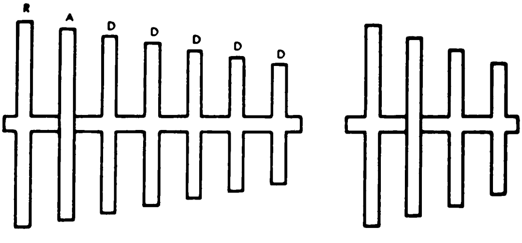
A parasitic element becomes a director when it is made shorter than the antenna element. In a director most of the energy is sent in a direction from the an-



(A) ANTENNA WITH REFLECTOR



(B) ANTENNA AND DIRECTOR



(C) 7 ELEMENTS YAGI (D) 4 ELEMENTS YAGI

Figure 7–4.—Parasitic array.

tenna element to the parasitic element. To see what takes place note the radiation pattern and the arrangement of the array in the diagram in (B) of figure 7–4. The director is usually about 5 percent shorter and about 0.1 wavelength from the antenna. Sometimes the impedance is reduced to 20 ohms at the driven element in this array.

Several parasitic elements can be used in conjunction with a driven antenna to increase further the directivity and power gain. The theoretical power gain of directional antenna arrays composed of an excited element and various numbers of parasitic elements is as follows:

Number of elements	*Power gain*
2	2.5 (4.0 db)
3	3.6 (5.6 db)
4	5.0 (7.0 db)
5	6.4 (8.1 db)

Notice the element array called a Yagi antenna in (C) of figure 7–4. In it the antenna or driven element is insulated, but the reflector and all directors are welded to a piece of tubing which runs parallel to the direction of propagation. The beam width of this array is about 19 degrees. The four-element Yagi shown in (D) is constructed similarly to the one in (C) and has a beam width of about 50 degrees.

The conventional method for describing the directivity of an antenna array is in terms either of the ratio of the power in the best lobe to the power radiated by a simple half-wave antenna (antenna gain), or the ratio of the power in the direction of the best lobe to the power in the opposite direction (front-to-back ratio). The direction of the best lobe is usually called FORWARD direction, while that in the opposite is called the BACKWARD direction. In other words, this ratio is the power in the forward field to the power in the backward field. For example, the front-to-back ratio of the antenna array illustrated in (B) of figure 7–4 is about 5:1. In decibels, this represents approximately a 7-db gain. The

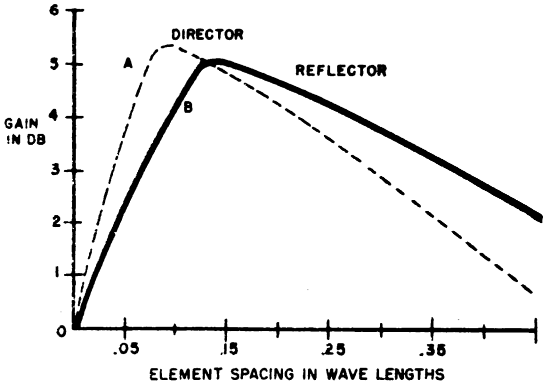


Figure 7–5.—Relative power gains of parasitic arrays for different reflector and director spacings.

ratio between forward power and power from a single half-wave antenna is about 4 db.

Figure 7–5 shows the effect of the element spacing on the power gain of an array as compared to the field strength of a half-wave antenna alone. Curve *A* shows the power gain for director spacings in wavelengths between a half-wave director and the driven half-wave radiator. Curve *B* shows the power gain for half-wave reflector spacings between the reflector and the driven half-wave radiator.

The graph shows that when a single parasitic element is used, there is little difference in the gain of the array if the parasitic element employed is used either as a reflector or as a director. When the parasitic element is tuned to work as a reflector, the spacing which gives maximum gain is about 0.15 wavelength and this maximum gain peak is fairly broad. The director will give slightly more gain than the reflector, but the difference between the two is less than 0.5 db. Consequently, on the basis of gain, there is little choice between the two types of operation. The broader curve of the reflector does offer a distinct advantage in that the spacing is less critical. Similarly, a small error in reflector spacing will reduce the gain of the array less than a same spacing error in a director type array. For this reason,

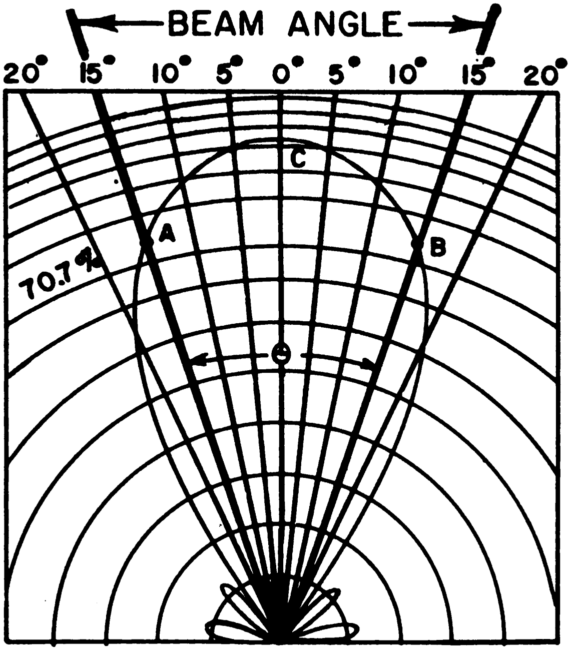
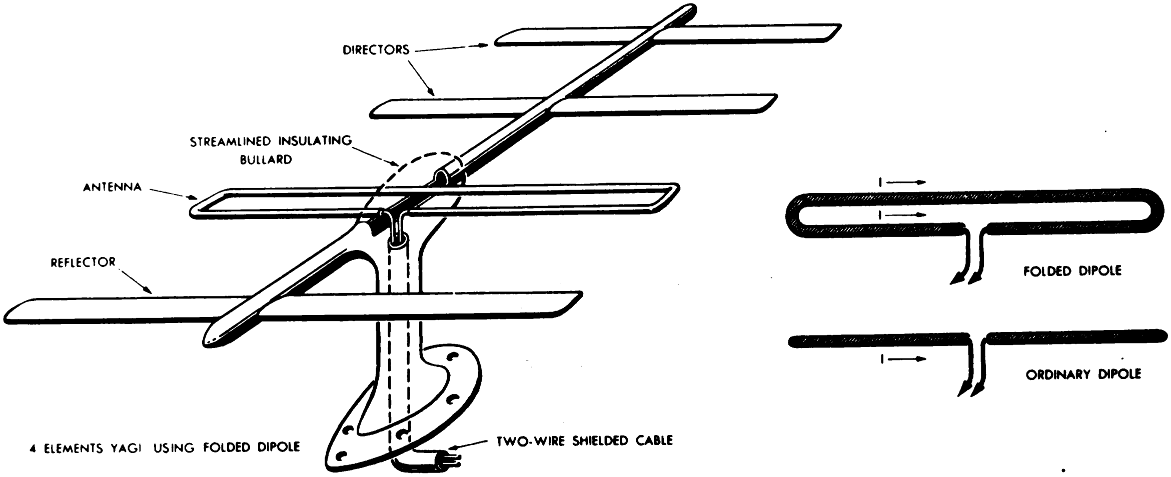


Figure 7–6.—Beam angle and half-power points.

when a single element is used, the array is usually a reflector type array.

Comparisons between the more directive types of arrays are made in terms of the beam angle. This angle is the angle between half-power points in the main lobe (radiation curve). In arrays the half-power points are the points where the electric field strength (volts per meter) is 0.707 as great as that along the axis of the beam. Figure 7–6 shows the beam angle for a radiation pattern. In this illustration the angle between points *A* and *B*, or the angle θ, is the beam angle. At each of these points the signal strength is 0.707 of its value at point *C*. At *A* and *B* the power radiated is one-half the value as at *C*, thus these points are also known as half-power points.

In multielement arrays the input impedance drops as low as 15 ohms. In these arrays special matching devices



Folded Dipole

Figure 7–7.—Folded dipole.

are needed for matching this low impedance to the higher impedance of most RF lines. One method for making the match is to use the Delta matching system in which the end of the transmission line that is to be connected to the antenna is spread apart (fanned out) before being connected. The section of the line that is spread apart has a continuously increasing characteristic impedance. The impedance of a half-wave radiator increases from 73 ohms in the center to about 2,500 ohms at the ends. Thus, the impedance at any point is determined by the distance from the center. Two points can be found on the radiator where the impedance equals that of the fanned section of the transmission line. The connections are made at these points. Another method, the one most often employed, uses the folded dipole as shown in figure 7–7.

A folded dipole is a full wavelength conductor which is folded to form a half-wave element. A better description is that it consists of a pair of half-wave elements connected together at the ends. In it the voltage at the ends of each element must be the same. In operation, the field from the driven element induces a current in the second element. This current is the SAME as the current in the driven element.

An ordinary dipole with a given current, I, produces a certain field intensity in space. Due to this field, there is also a certain power density per square meter in space. This power density is produced by the input power, P. The relationship between the input resistance, the current, and the input power is expressed by the equation,

$$R=\frac{P}{I^2}$$

With the folded dipole, the same current, I, exists at the input terminals; but this same current flows in both sections of the antenna, and consequently, the field strength in space is doubled. This causes the power density per square meter to increase to four times the previous value when the CURRENT AT THE INPUT REMAINS UNCHANGED. In turn the input power must be four times as great. In this case, to balance the equation, it is necessary to multiply R by 4 as follows:

$$\frac{4P}{I^2} = 4R$$

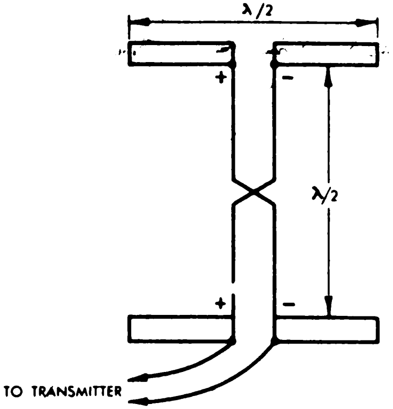
So long as each section of a dipole has the same diameter, the input resistance is four times that of the simple half-wave dipole. Increasing the diameter of one section makes the increase in impedance still greater. The input impedance to the driven element of a parasitic array drops to about a fourth of the value of the coaxial cable impedance, but the use of a folded dipole increases the impedance by about four times. In this way a good impedance match is effected.

Interconnection of Elements in Arrays

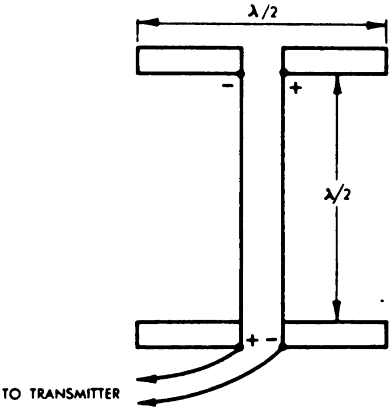
There are a variety of ways to connect the elements in an antenna array to obtain the required phase of excitation. The most convenient method to change the phase from one element to another element when they are a half wave apart is with an RF line. In the broadside array shown in (A) of figure 7–8, one element is excited directly from the transmitter. This element is connected to the other element by a half wavelength of RF line. The phase of the voltage along the RF line is shifted 180 degrees per each half wavelength. The leads to the second driven element are reversed. This causes a reversal of polarity similar to that resulting from another 180-degree phase shift. The second element is then excited in the same phase as the first.

It is possible to connect any number of elements in this manner. In addition, increasing the number of elements makes the array more directional.

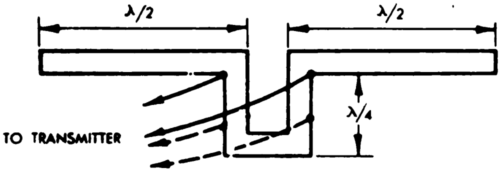
In the end-fire array in (B) of figure 7–8, the elements are also interconnected with RF lines. But instead of reversing the leads to the elements, they are connected directly to take advantage of the phase shift. With half-



(A) BROADSIDE ARRAY



(B) END FIRE ARRAY



(C) COLLINEAR ARRAY

Figure 7–8.—Connecting half-wave elements to obtain proper phase relationship.

wave spacing, the interconnecting line will be a half wave long and provide a 180-degree phase shift. If quarter-wave spacing is used, the interconnecting line will be a quarter wave long. This arrangement produces a 90-degree shift.

The usual method for providing the correct phase in a collinear array is shown in (C) of figure 7–8. Since the current direction changes for each half wavelength of an antenna, it is not possible to connect the half-wave sections together directly. Instead, the half-wave section which carries the current in the wrong direction is folded to form a quarter-wave section of RF line. This brings together the ends of the sections in which current flows in the same direction. In other words, in terms of voltage it is necessary that the two antenna sections have voltages which are opposite polarity at the open end. Further, note how the line to the transmitter is connected in (C). The connection is made at a point of very high impedance. This requires the use of a resonant line. If, however, a nonresonant line is to be used, it must be connected as shown by the dotted lines. In the diagram the connections near the closed end of the quarter-wave section is a point of low impedance.

The Parabolic Reflector

When a multielement broadside array is excited, the E-field which exists in front of the antenna will be in a single plane, as shown in (A) of figure 7–9, rather than in an arc as is the case in a single half-wave element. The larger the dimensions of this plane in terms of wavelengths, the greater is the directivity of the antenna system, and the narrower the radiated beam.

Although a multielement broadside array gives good results, it is quite complicated in structure. In it every element must be driven and all spacings and dimensions must be quite exact. A much simpler device for producing an electric field in a single plane is the parabolic reflector. As shown in (B) of figure 7–9, the parabola has its focal point at *F*. If a single antenna is placed

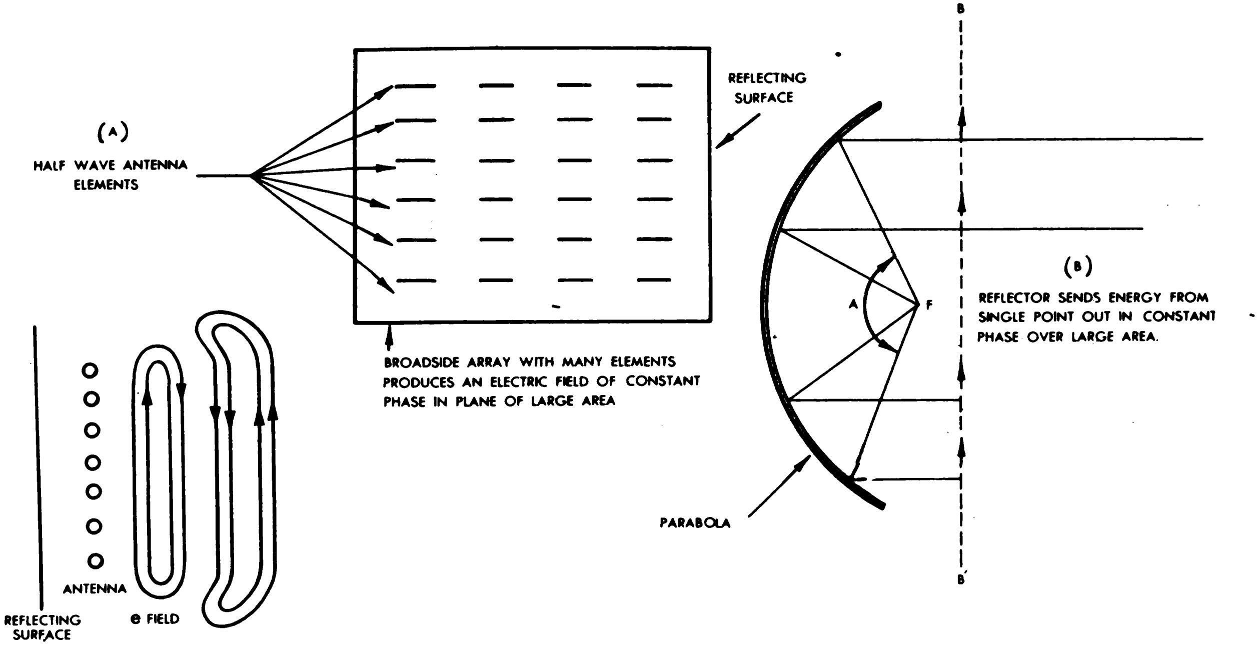

(A)
HALF WAVE ANTENNA ELEMENTS
REFLECTING SURFACE
BROADSIDE ARRAY WITH MANY ELEMENTS PRODUCES AN ELECTRIC FIELD OF CONSTANT PHASE IN PLANE OF LARGE AREA
ANTENNA
REFLECTING SURFACE
e FIELD
B
B′
A
F
PARABOLA
(B)
REFLECTOR SENDS ENERGY FROM SINGLE POINT OUT IN CONSTANT PHASE OVER LARGE AREA.

Figure 7–9.—Directional radiation by means of reflectors.

at *F* and caused to radiate a field, the electric field will leave the antenna in all directions at the same rate in the form of an arc as indicated at point *A*. As each part of the wavefront reaches the reflecting surface, it is shifted 180 degrees in phase and sent outward at an angle of reflection that is equal to the angle of incidence. All parts of the field will arrive at line *BB′* at the same time after reflection because all paths from *F*, to the reflector, to line *BB′* are equal in length. Thus, you see that with only one antenna and a specially shaped reflector, it is possible to produce a large electric field in a single plane. Or, looking at it another way, all parts of the field travel in parallel paths after reflection from the parabola in such a way that the rays are focused like the headlight beam from an automobile.

Like the broadside array, high directivity is not obtained until the diameter of the parabolic reflector is made many wavelengths long. This prohibits the use of the parabolic reflector at low frequencies, but for three-centimeter and ten-centimeter radar equipment these reflectors are very practical.

In figure 7–10, showing the exciting antenna for a parabolic reflector, the half-wave dipole is mounted a quarter wave back from the short on the coaxial line. To sharpen the focal point the antenna is physically less than a half wave long. However, the balls at the end make it electrically a half wave long. This broadens the band of frequencies it will handle. The airtight cylinder in which the antenna is enclosed permits the coaxial line to be pressurized. The inner surface of half the cylinder, that is, the side away from the parabolic reflector, is coated with a reflecting foil. This reflecting surface directs energy from that side of the antenna into the large reflector. Without this reflector half of the antenna radiation would be nondirectional.

The natural directivity of a dipole causes the pattern from a parabolic reflector to be somewhat sharper in the plane containing the dipole than in the other plane. For

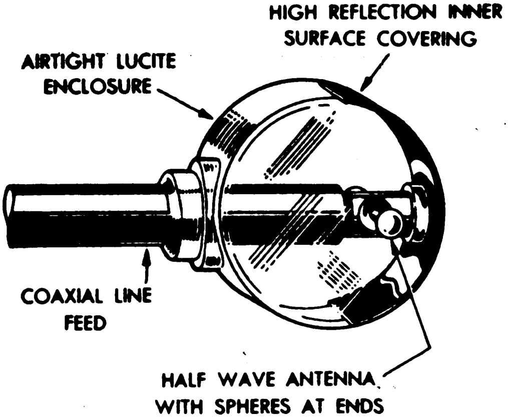

HIGH REFLECTION INNER SURFACE COVERING
AIRTIGHT LUCITE ENCLOSURE
COAXIAL LINE FEED
HALF WAVE ANTENNA WITH SPHERES AT ENDS

Figure 7–10.—Dipole for exciting parabolic reflector in 10-cm. system.

this reason, the dipole is erected horizontally for maximum azimuth accuracy in radar systems. If vertical accuracy is of primary importance, the dipole is mounted vertically.

Other methods of preventing the direct forward radiation use a parasitic reflector with the half-wave dipole, and a disk which is placed in front of the dipole.

When a waveguide is used in the RF system, it is possible to send the energy into the parabolic reflector with a horn radiator. In this case an orangepeel reflector, or section of a complete circular paraboloid may be used as the reflector. The feed is by a waveguide and a horn-type radiator. The horn-type radiator and orange-peel parabola are constructed in such manner that the radiation pattern of the horn is such that almost all of the energy that it radiates is intercepted and reflected by the reflector. This arrangement is highly directive in the vertical plane and is used principally to determine the altitude of airplanes.

COSECANT-SQUARED REFLECTORS.—In airborne radar sets that are designed for use in scanning a large area, a special beam shape is required. This is shown in figure 7–11. The beam should be narrow in the horizontal plane but broad in the vertical plane.

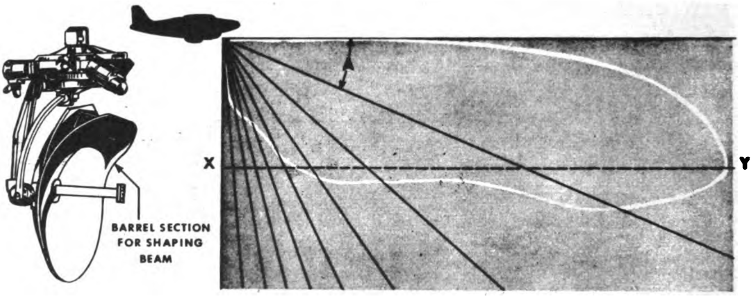


Figure 7–11.—Cosecant-squared antenna and pattern.

In addition, the particular section of the area that is being scanned must not be uniformly illuminated by the radar beam. Instead, the objects near the aircraft should be illuminated by a relatively smaller amount of power than those farther away in the same sector. This is required because echo signals from nearby objects will be strong due to their closeness to the plane. On the other hand, a greater amount of power should fall on more distant objects in this sector since they would normally produce the weakest echoes. Thus, the radiated field produced by the beam should increase in intensity at greater distances. Or, in other words, the field intensity should vary as the cosecant of the angle between the horizontal line of the plane and a line drawn between the aircraft and a given point on the ground. Since power varies as the square of the field intensity, then the required power pattern must be a cosecant-squared pattern.

A common method used to obtain such a pattern is to use a special parabolic reflector in which the top portion is bent forward in order to produce the required pattern. Such a reflector is shown in figure 7–11 along with the radiation pattern produced.

Another type of reflector is shown in figure 7–12. This reflector differs from the one just discussed in that it is capable of producing either a pencil type (narrow beam both vertically and horizontally) or a cosecant-squared pattern. Mounted on the reflector is a rod grating, rough-

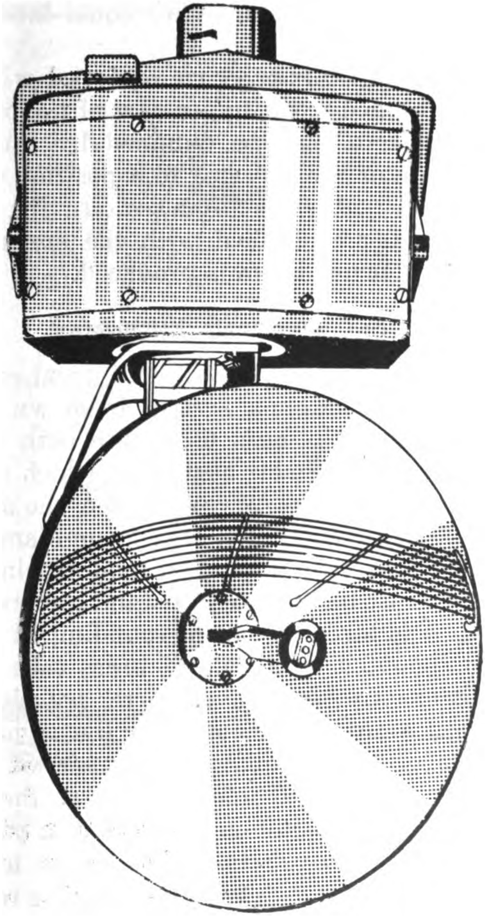


Figure 7–12.—Parabolic antenna with spoiler.

ly shaped in the form of a barrel stave, which is polarized and acts as a spoiler to deform the pencil-beam pattern into an equal-energy-return (cosecant-squared) pattern. The choice of pattern desired may be obtained by rotating the entire reflector and grating 90 degrees by means of a beam-pattern-change motor and gear train so that, in the position where the spoiler is nonreflective, a pencil-beam pattern is produced, and, in the position where the spoiler is reflective, a narrow, blade-like, equal-energy-return pattern results.

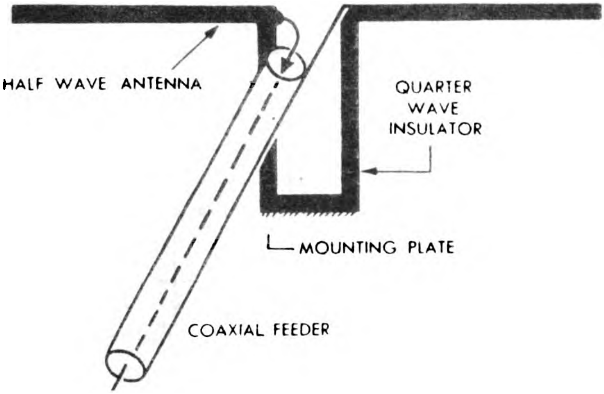
Some reflectors of this type are equipped with a solid spoiler that is hinged to the upper part of the paraboloid. The desired pattern may be obtained by causing this spoiler to tilt out of the reflector to a position where the cosecant-squared pattern is produced. In the retracted position the spoiler is flushed with the surface of the paraboloid and a pencil beam is produced.

TYPICAL AIRBORNE ANTENNAS

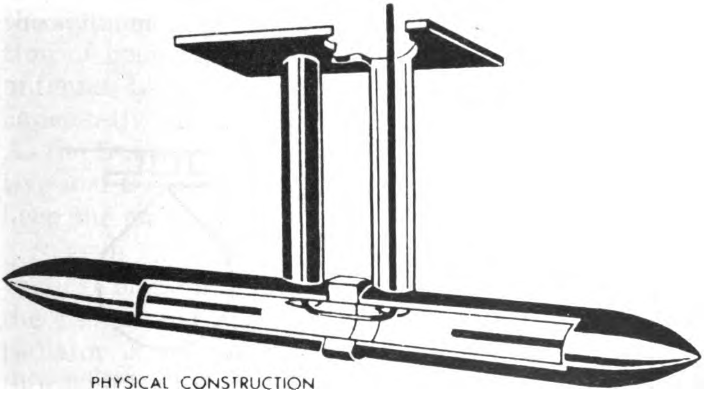
A UHF BROAD-BAND ANTENNA.—A typical example of a directional UHF antenna system is shown in figure 7–13. This antenna has been used with both frequency modulation and pulsed radar altimeters which required a wide-band characteristic. As you can see, the antenna is a half-wave dipole constructed of a large diameter tube which is divided at the middle and mounted in the slipstream. As it must be able to withstand air pressures at high speeds, it is streamlined by rounding at the ends. It is rigidly mounted on a pair of brass tube supports. Each tube is exactly equal to a quarter wavelength and is welded to a plate that mounts on the airplane. Together, the tubes form a quarter-wave section which is shorted at the mount end and is open at the antenna end. Each half of the antenna connects to a part of the leg of this quarter-wave section. The entire antenna is perfectly insulated by a quarter-wave metallic insulator.

The entire section is connected electrically to the antenna through a coaxial cable. One of the tubes supporting the antenna is the outer conductor for the coaxial cable. A wire through the center of this tube starts from a coaxial connector at the mounting plate and ends at the opposite half of the antenna. Because of the split sections, a bakelite rod inside the antenna tubing is used to give the antenna mechanical strength. The half sections are insulated from one another by a ceramic bushing.

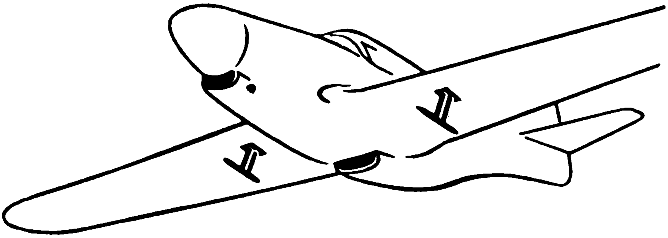
In an installation of this type two antennas are em-



EQUIVALENT CIRCUIT

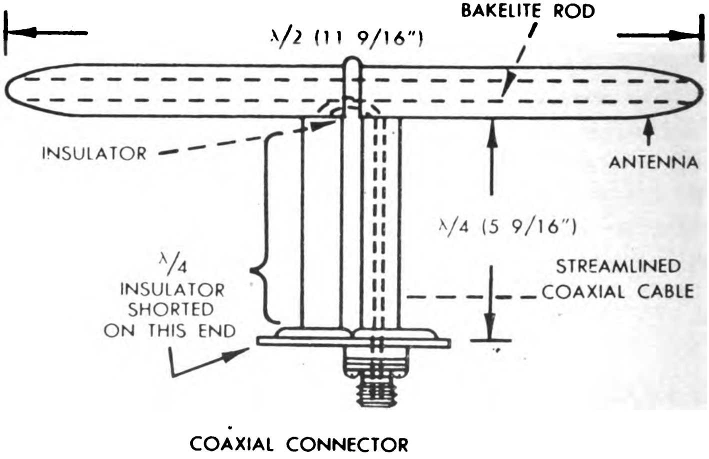


PHYSICAL CONSTRUCTION



MOUNTING OF 400 MC ANTENNAS ON PURSUIT PLANE

Figure 7–13.—A UHF antenna.



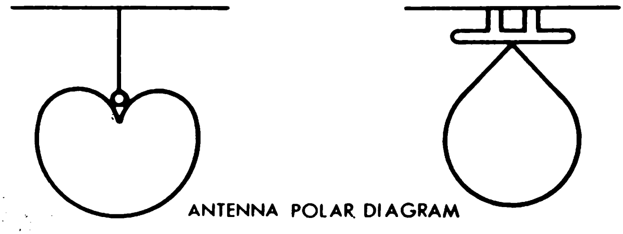


Figure 7–14.—400-mc. antenna dimensions and polar diagrams.

ployed—one for transmitting and the other for receiving. Usually each is mounted under a wing. Usually the part of the aircraft which is located between the antennas serves to attenuate any direct signal radiation between them. If this arrangement is not possible, the antennas are mounted end to end along the bottom of the fuselage. Since a half-wave dipole has very little radiation off the ends, the receiving antenna is still isolated from the transmitting antenna. The signal desired is the one that goes from the transmitting antenna to the earth's surface and then returns to the receiving antenna. Each antenna is mounted a quarter wavelength from the metal skin of the airplane. The image antenna in the skin is a half wave from the real antenna and parasitically excited out of phase. This produces a virtual two-element endfire array. Since no radiation fields go through the skin, all radiation is directed downward in a somewhat narrowed lobe, as you can see in the side and front view of the radiation pattern shown in figure 7–14. The large diameter of the antenna itself makes it broadly resonant and causes its characteristics to be fairly uniform over the 40-mc. bandwidth of the equipment.

Another factor that adds to its uniform response over a wide band is its method of support; this support is a shorted quarter-wave section of transmission line. Since the antenna acts the same as an open quarter-wave section of transmission line, as the frequency increases the antenna becomes inductive and the support will be capacitative, thus, tending to keep the antenna resonant. As the frequency decreases the antenna becomes capacitative and the support will be inductive; this also tends to keep the antenna resonant.

SLEEVE ANTENNA.—The sleeve or coaxial antenna is a vertical omnidirectional antenna developed to minimize the radiation from the feed line. One type of coaxial radiator is one-half wavelength long. Its top half is a thin whip or rod, and the bottom half is a comparatively large hollow metal tube or cylinder. The coaxial feed line is run up through the sleeve radiator or cylinder, as shown in (A) of figure 7–15. The center conductor of the coaxial line connects to the whip radiator and the outer conductor connects to the sleeve radiator. Since there is no RF field within the sleeve, interaction between the radiator and the coaxial line is minimized.

The section of transmission line within the sleeve acts in conjunction with the sleeve to form a short-circuited concentric line which has an extremely high impedance across the open end. This high-Q parallel-resonant circuit minimizes coupling between the sleeve radiator and the

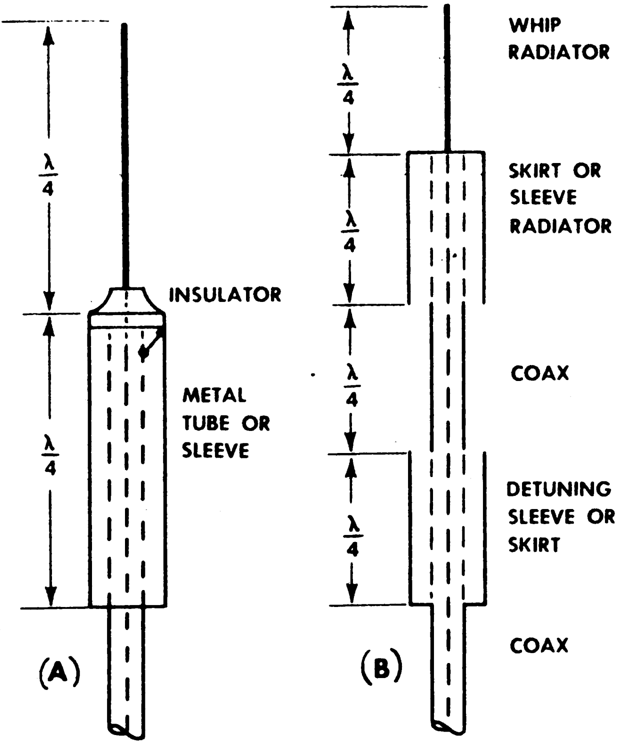


Figure 7–15.—Sleeve antenna.

transmission line or hollow supporting mast. Although the sleeve effectively decouples the section of line within it from the sleeve radiator itself as well as from the upper radiator, there will still be some coupling between radiators and the section of the transmission line or mast immediately below the sleeve. This coupling causes some radiation from the vertical portions of the line and/or mast and, as a result, some power is wasted in high-angle radiation. Radiation can be eliminated from the feeder and mast by placing a quarter-wave detuning sleeve below the sleeve radiator as shown in (B) of figure 7–15. The detuning sleeve is mounted with its open end facing upward, one-quarter wavelength below the bottom of the sleeve radiator. Its closed end is connected to the mast, or to the outer conductor of the coaxial feed line.

Inasmuch as the coaxial antenna is effectively a vertical, center-fed, half-wave radiator, its center impedance is approximately 70 ohms. A 70-ohm feed line provides the most efficient match, but lines up to 120 ohms can be used without serious mismatch. It is possible to maintain a uniform impedance over a larger frequency band by increasing the diameter of the dipole.

The optimum sleeve length usually runs from 0.98 to 1.04 of a free-space quarter wavelength. The optimum length for the whip radiator varies from about 0.94 to 0.96 of a quarter wavelength.

GROUND-PLANE ANTENNA.—A vertical quarter-wave antenna several wavelengths above ground produces a high angle of radiation which is very undesirable at the very high and ultrahigh frequencies. The most common means of producing a low angle of radiation from such an antenna is to work the radiator against a simulated ground called a GROUND PLANE. A simulated ground may be made from a large metal sheet or several wires or rods radiating from the base of the radiator. An antenna so constructed is known as a GROUND-PLANE ANTENNA.

Several types of ground-plane antennas are shown in figure 7–16. The radius of the ground sheet or the length of the rods used should be one-quarter wavelength or longer. The function of a ground plane corresponds to that of a counterpoise used with low-frequency vertical antennas.

This type antenna is frequently used with VHF and UHF communications equipment in aircraft. The skin of the aircraft serves as the ground plane. When this type antenna is used with a ground installation, it is necessary to employ a ground plane as illustrated in figure 7–16. Without the use of a ground plane it will be impossible to obtain proper loading and efficient radiation.

Since there are no ordinary transmission lines having

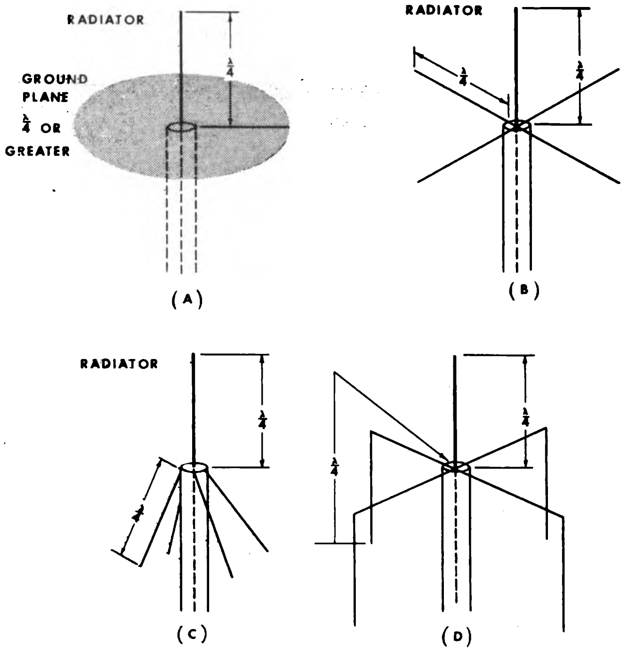


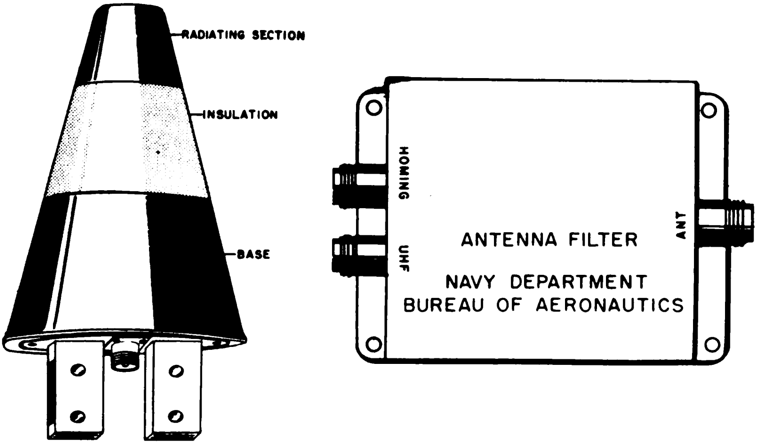
Figure 7–16.—Ground-plane antennas.

a characteristic impedance low enough to permit a direct connection to the 20- to 36-ohm impedance of the ground-plane antenna, it is necessary to use a quarter-wave matching section to obtain an acceptable match.

If it is desired to avoid the use of a matching section with its accompanying power losses and bandwidth limitations, the radials can be arranged to slope downward in the manner of an inverted cone as shown in (C) of figure 7–16. The impedance increases as the angle between the radials and the horizontal plane increases. This method can be used to secure a suitable match to a 52-ohm transmission line. The impedance can be increased further to match a 72-ohm line by bending the radials down parallel to the transmission line or supporting shaft as shown in (D).

An ultrahigh frequency, streamlined blade, broad-band antenna for general-purpose use over the range of 225 to 400 megacycles per second is shown in figure 7–17. This antenna consists essentially of a near quarter-wave stub, fed through a matching section within the base. It is designed for mounting upon the exterior skin of an airplane. The antenna is streamlined for high speeds. Its cross sections take the form of ellipses on the leading half of the body and of parabolas on the trailing half. The resulting form has a minimum drag effect. The antenna is designed to withstand pressures resulting from airspeeds up to 600 miles per hour.

The antenna is one self-contained unit, made up of three sections all securely fastened together. The base section and radiating tip are made of aluminum, the central section is a molded resin reinforced with glass fibers. No accessories are required other than those for mounting and connecting. The radiating element is a solid aluminum casting with an inner cone-shaped nipple curving down to the feed point.



(A) (B)

Figure 7–17.—UHF blade antenna.

A single broad-band antenna of this type may be connected to two or more pieces of equipment that operate on different frequencies. The limiting factor is that the equipments must operate within the frequency range of the antenna. When this is done it is necessary to use band-pass filters in series with the antenna cables. These filters provide the attenuation that is necessary in order to isolate the various equipments that are connected to the single antenna. A filter of this type is shown in figure 7–17 (B).

These filters consist of lumped constants, coils, and capacitors and are designed to have 1 db or less insertion loss in the pass bands. The attenuation in the bands varies from 25 db to over 80 db depending on the frequency. The power handling capacity of each filter is usually over 700 watts peak power in the pass band, the average power capacity being about 100 watts depending on the frequency.

The antenna shown in (A) of figure 7–17 could be used with a UHF transceiver and a UHF homing receiver at the same time.

SLOT ANTENNA.—The slot antenna consists of a radiator formed by cutting a narrow slot in a large metal surface. Such an antenna is shown in figure 7–18. The slot length is a half wave while the width is a small fraction of a wavelength. Such an antenna is frequently compared to a conventional half-wave dipole consisting of two flat metal strips whose sizes are such that they would just fit into the slot cut out of the large metal sheet. This comparison is made since the radiation pattern produced by the slot antenna when cut into an infinitely large metal sheet and that of the complementary dipole antenna are the same.

It is of interest to note two important differences between the slot antenna and its complementary antenna. First, the electric and magnetic fields are interchanged. In the case of the dipole antenna shown, the electric lines are horizontal while the magnetic lines form loops in the

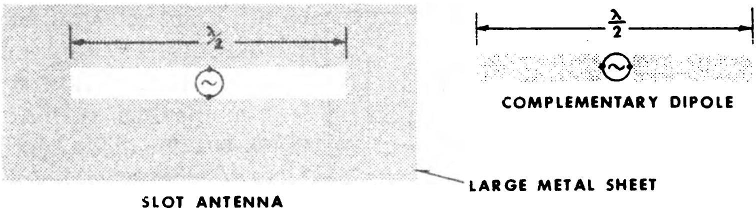


Figure 7–18.—Slot and complementary dipole antenna.

vertical plane. With the slot antenna, the magnetic lines are horizontal and the electric lines are vertical. These electric lines are built up across the narrow dimension of the slot. As a result, the polarization of the radiation produced by a horizontal slot is vertical. If a vertical slot is used, the polarization is horizontal.

A second difference between the slot antenna and its complementary dipole is that the direction of the lines of electric and magnetic force reverse abruptly from one side of the metal sheet to the other. In the case of the dipole, the electric lines have the same general direction while the magnetic lines form continuous closed loops.

When energy is applied to the slot antenna, currents flow in the metal sheet. These currents are not confined to the edges of the slot but rather spread out over the sheet. Radiation then takes place from both sides of the sheet.

A coaxial line is frequently used to feed the slot antenna. The outer conductor of the feed line is bonded to the metal sheet while the inner conductor is connected to the opposite side of the slot as shown in figure 7–19. Note that the coaxial line is not connected at the center of the slot length since a severe impedance mismatch would occur.

The impedance at the center of a slot whose electrical length is a half wave is 530 ohms. This impedance is reduced to a lower value as the input connection is moved toward the end of the slot. At a point about one-twentieth of a wavelength from either end, the input

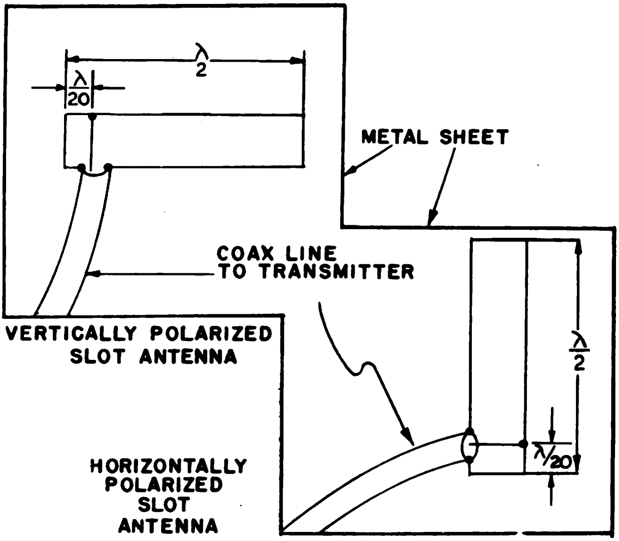


Figure 7–19.—Feeding slot antennas.

impedance falls to about 50 ohms. A coaxial line may then be connected here without serious mismatch. The slot length may be increased to a full wave if it is desired to connect the coaxial line to the center of the slot. With a full-wave slot, the input impedance at the center of the slot is 50 ohms.

Frequently, it is desired to produce radiation from only one side of the metal sheet in which the slot is cut. This is conveniently arranged by boxing in the slot with a section of waveguide. If the depth of the waveguide is a quarter wave (fig. 7–20), then the waveguide will introduce no reactance. Under these conditions, the

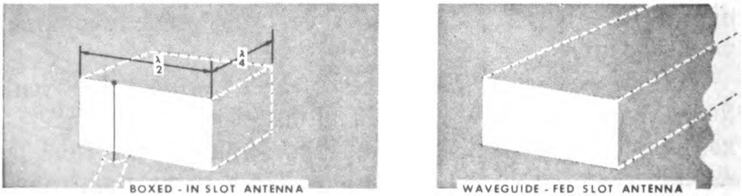


Figure 7–20.—Unidirectional slot antennas.

input impedance at the center of the slot is about 1,000 ohms and the impedance at a point about one-twentieth of a wavelength from either end is about 100 ohms.

Another convenient method of making a slot antenna radiate from only one side of the metal sheet in which it is cut is to energize the slot by means of a waveguide as shown in figure 7–20. This does away with the coaxial feeder altogether.

It has been mentioned that the radiation pattern produced by a slot antenna is the same as that produced by a complementary dipole antenna. This is true only if an infinitely large metal sheet is used. As the width of the metal sheet is reduced, the amount of radiation that occurs at small angles to the plane of the metal sheet is reduced considerably. This is shown in figure 7–21. The radiation pattern produced by a complementary dipole antenna in the plane perpendicular to the antenna is also shown for comparison. In this pattern, you are looking at the end of the dipole. If all radiation disappeared to the left of the vertical centerline shown in the illustration, then you would have the radiation pattern produced by a boxed-in slot antenna. (See (A) of fig. 7–21.) Here you are looking along the edge of the metal sheet in which the slot is cut. The length of the slot is at right angles to the plane of the page and the radiation patterns shown are in the plane of the polarization of the energy.

If the width of the metal sheet is a little over 5 wavelengths, then the scalloped pattern shown in (B) results. As the width of the sheet is further reduced to slightly over 2½ wavelengths, fewer irregularities appear in the pattern, as shown in (C). Also note the reduced radiation at a fixed angle (say 30 degrees) with respect to the metal surface. At a width of only about a half wavelength, the pattern shown in (D) bears only a little resemblance to that shown in (A). The radiation produced at an angle of 30 degrees is further reduced. In practice, since the size of the metal surface in which the slot is cut is usually quite large (compared to a wave-

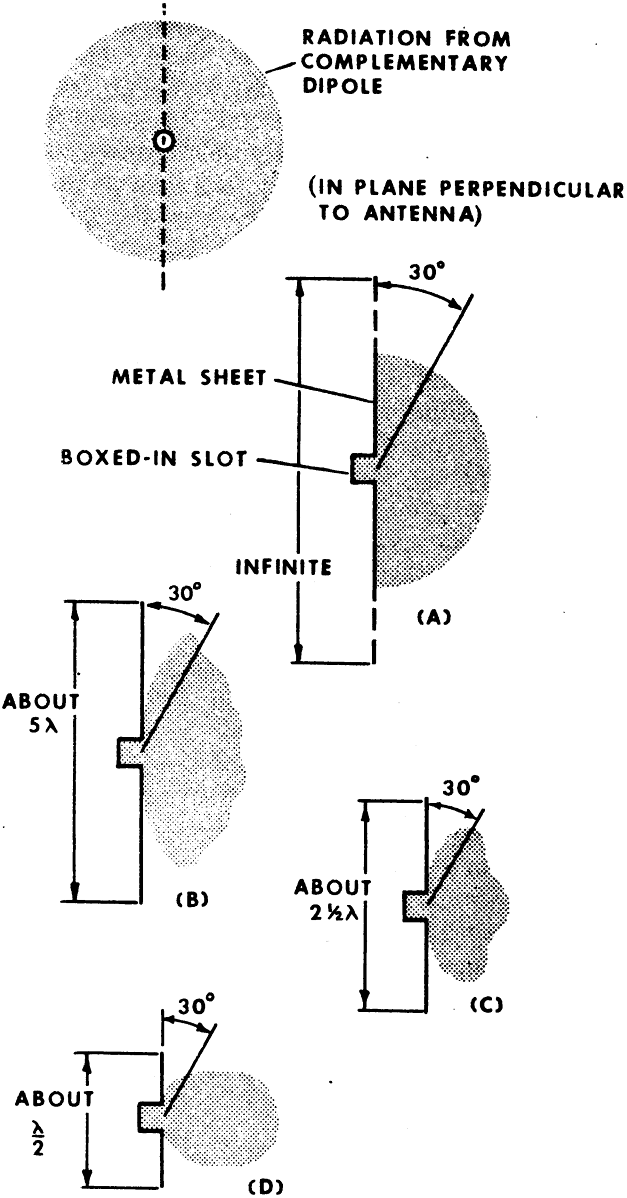


Figure 7–21.—Radiation from boxed-in slot antennas.

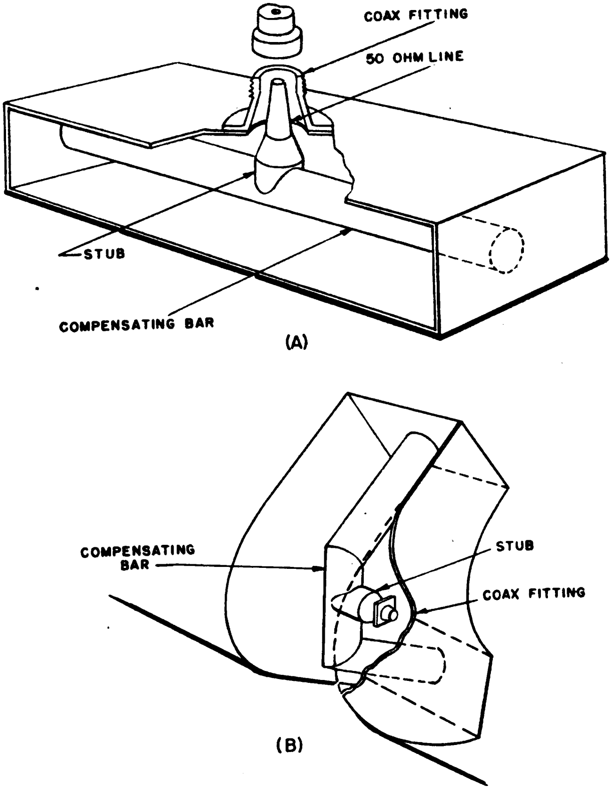


Figure 7–22.—T-fed slot antennas.

length), radiation patterns having a form somewhere between that shown in (A) and (B) are common.

A type of slot antenna which was developed for use over a very wide frequency band and which could be fed by coaxial line is shown in figure 7–22. With the compensating bar (providing capacitive loading) and stub as shown, this antenna has a standing-wave ratio of less than 2 to 1 over a frequency range from about 30 percent below to about 30 percent above the design center frequency when fed with 50-ohm coaxial cable. Such an antenna may be constructed in a large flat metal surface or it may be made to conform with a curved metal surface. For example, the curved slot antenna shown in (B) can be installed in the leading edge or tip of an aircraft wing. With such a severely curved large metal surface, the radiation pattern produced is quite different from that produced with a slot in a flat surface. The extent of the curvature has considerable effect on the radiation produced.

Still another form of the slot antenna is shown in figure 7–23. Here the large metal sheet is eliminated completely. Instead, an array of diagonal slots is cut into a length of waveguide. The slots are spaced at intervals along the waveguide so that they are excited in phase. Maximum radiation takes place in the direction indicated.

Slot antennas are adaptable for the VHF and UHF ranges. Their most important practical advantage is that the feed section which energizes the slot may be placed below the large metal surface in which the slot is cut. Thus, nothing need protrude from the surface. In addition, the slot itself may be covered by a section of insulating material so that the opening is closed. As a result, no air resistance occurs. This is particularly important in the case of high-speed aircraft where the vibration problems and drag that would be introduced by conventional antennas cannot be tolerated.

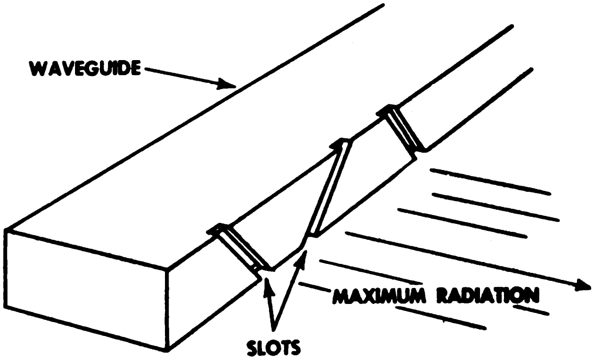


Figure 7–23.—Slot array.

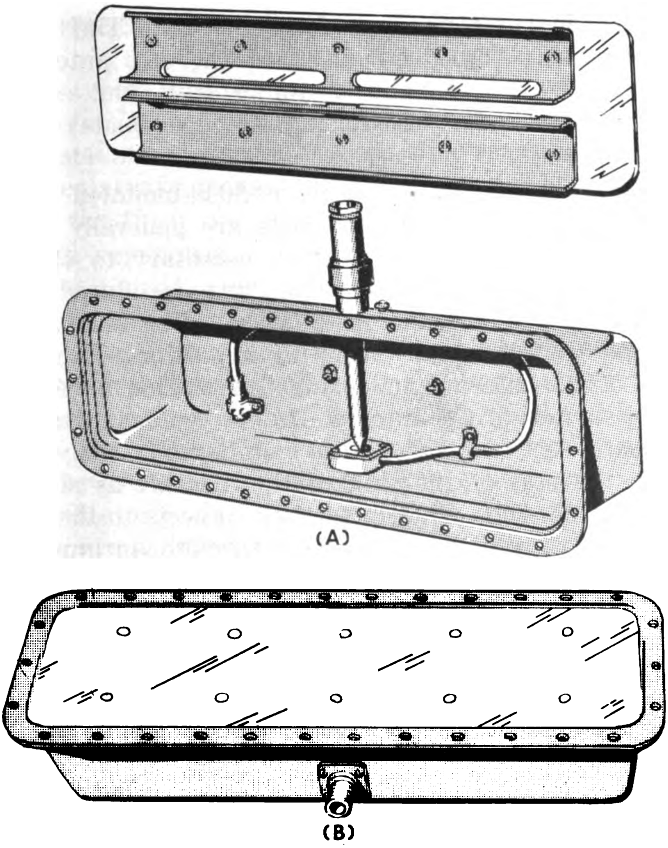


Figure 7–24.—Radio altimeter slot antenna.

When a slot antenna is installed on an aircraft, particular attention should be paid to the slot location with respect to large protuberances. Wings, fuselage, or engine will act to cut off a large portion of the energy radiated from a slot if these objects are located between the slot and the receiving location. Favorite locations for slot antennas on aircraft include the leading edge of the wing and the side of the nose.

The slot antenna illustrated in figure 7–24 is used as the radiator with airborne radio altimeters. Two such flush-mounted antennas are used in the installation, one for transmitting and one for receiving. The internal arrangement (fig. 7–24 (A)) of the slot antenna includes the feed element, a tuning stub, and capacitor plates (attached to the fiber-glass cover plate) to load the cavity.

A SKIN ANTENNA is a type of flush-mounted antenna used in aircraft. Such antennas are generally used in high-speed aircraft where it is necessary to eliminate the speed-reducing aerodynamic drag produced by the usual types of antenna installations. The skin antenna is made by electrically isolating a portion of the metal skin of the aircraft and using this section as the radiating surface of the antenna. In one type of installation, a plastic section insulates the metal skin at the wingtips from the rest of the skin of the aircraft as shown in figure 7–25. The wingtip section connects to the receiving and transmitting equipment through an impedance-matching network and coaxial cable.

The range of a skin antenna approximates that of fixed-wire antennas for the same frequency coverage. Although the illustration in figure 7–25 shows a wingtip installation, skin antennas can also be made a part of suitable tail or fuselage surfaces. The radiation patterns depend on the operating frequency, the size of the radi-

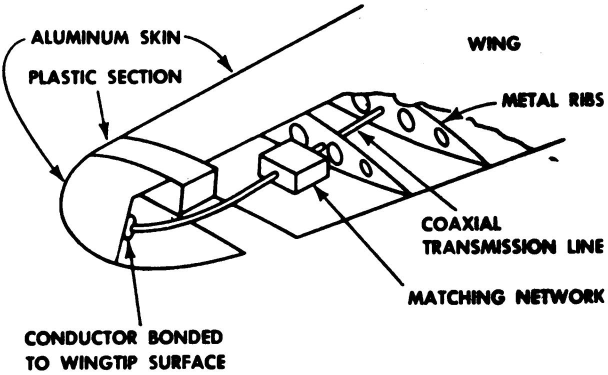

ALUMINUM SKIN
PLASTIC SECTION
WING
METAL RIBS
COAXIAL TRANSMISSION LINE
MATCHING NETWORK
CONDUCTOR BONDED TO WINGTIP SURFACE

Figure 7–25.—Wingtip antenna.

ating surface, and the location of that surface on the aircraft.

Other types of flush-mounted antennas which find wide use in VHF and UHF aircraft communications and radar installations consist of loops, dipoles, and parasitic and driven arrays encased in plastic housings or canopies forming the tip of a wing, elevator, stabilizer, or nose.

CORNER REFLECTOR.—When a unidirectional radiation pattern is desired, such a characteristic can be obtained by the use of a corner reflector. A corner-reflector antenna is a half-wave radiator with a reflector consisting of two plane metal surfaces meeting at an angle immediately behind the radiator. In other words, the radiator is set in the plane of a line bisecting the corner angle formed by the reflector sheets. The construction of a corner reflector is shown in figure 7–26. Corner-reflector antennas are mounted with the radiator and the reflector horizontal when horizontal polarization is desired. In such cases, the radiation pattern is very narrow in the vertical plane, with maximum signal being radiated in line with the bisector of the corner angle. The directivity

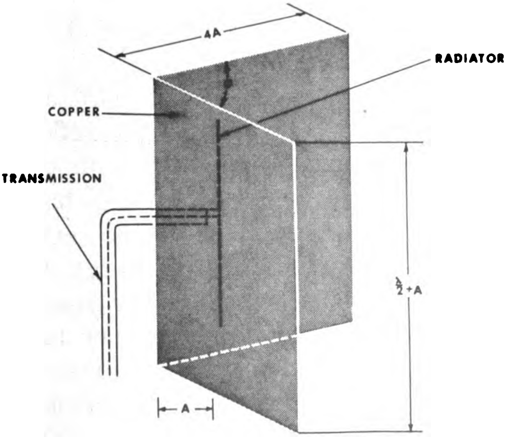


Figure 7–26.—Corner-reflector antenna.

in the horizontal plane is approximately the same as for any half-wave radiator having a single-rod type reflector behind it. If the antenna is mounted with the radiator and corner reflector vertical (as shown), maximum radiation is produced in a very narrow horizontal beam. Radiation in a vertical plane will be the same as for a similar radiator with a single-rod type reflector behind it.

At frequencies in the VHF and UHF bands, the corner reflector is often made from solid metal sheets. At lower frequencies, such sheets are massive and are easily damaged by high winds. Therefore, screen wire or metal rods located in the plane of the reflector sheets and running parallel to the radiator are used instead. The spacing between the rods must not be more than 0.06 wavelength at the operating frequency.

The gain and the radiation resistance of a corner-reflector antenna are both affected by the reflector corner angle Φ and the spacing, A, between the radiator and the corner as shown in figure 7–26, as well as the construction of the antenna and reflector. The gain (expressed as a power ratio between the dipole with a reflector and the dipole alone) for several corner-reflector antennas is shown in the graph in figure 7–27. Here the gain is shown for various spacings between the dipole and the corner reflector, and for an angle Φ of 90, 60, and 45 degrees.

Sometimes the corner reflector is used as a PASSIVE RADIATOR. Under these conditions, the reflector is NOT excited by a dipole that is mounted with the reflector. Instead, the reflector is acted on by a distant transmitting antenna. Here, the purpose of the corner reflector is to return the greatest amount of energy possible back in the direction from which it originally came.

The most important use of the corner reflector as a passive radiator is in radar applications. Here, it is sometimes necessary to make good radar targets from objects which do not ordinarily reflect back enough of the transmitted radar beam to disclose their presence.

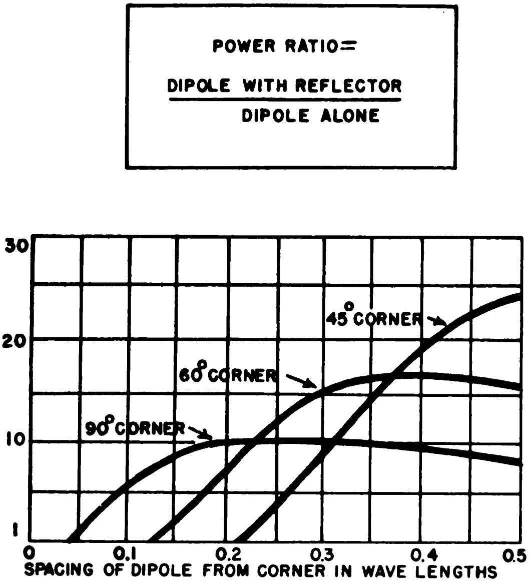


Figure 7–27.—Gain of corner-reflector antenna.

For example, it is normally very difficult for a searching aircraft to locate by its radar a rubber liferaft, particularly in a rough sea. If a corner reflector is installed on the raft, then it becomes possible for an aircraft radar to detect the raft over a distance of many miles. Corner reflectors are also used to mark the edges of landing fields or a particular runway so as to provide a well-marked point for the aircraft or ground radar equipment. Corner reflectors are frequently installed atop important buoys or strategic points of land which would otherwise not be good radar targets.

The appearance of a corner reflector used for marking radar targets is shown in figure 7–28. Three mutually perpendicular metal surfaces similar to the arrangement of the floor and two side walls at a room corner are used. It should be noted that in the passive corner reflector the corner angle Φ is always 90 degrees. If a radar beam is directed into the reflector, reflections may occur from all three surfaces (as shown in the illustration), from

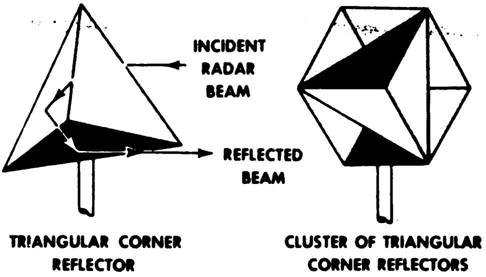

INCIDENT RADAR BEAM
REFLECTED BEAM
TRIANGULAR CORNER REFLECTOR
CLUSTER OF TRIANGULAR CORNER REFLECTORS

Figure 7–28.—Passive corner reflectors.

two surfaces, or from one surface, depending on the angle of arrival of the beam. In any case, the reflections cause the radar beam to be returned in the direction from which it came. Under these conditions, the radar antenna picks up the reflected energy and detection occurs.

A single corner is effective only for energy that arrives from directions covering one octant of a sphere centered at the reflector. If it is desired to have coverage in all directions (such as would be the case for liferaft detection), then a cluster of eight such corner reflectors is used.

The size of such reflectors is not critical so long as the length of a short side is many wavelengths long at the frequency of the detecting radar. The corner-reflector cluster which is included as part of the emergency rescue equipment on liferafts is a collapsible structure made of screen and measuring two feet on a short side. The reflector is mounted about four feet above the raft.

CONE ANTENNA.—It frequently becomes desirable for a transmitting or receiving antenna to operate over a wide band of frequencies. This is especially true in electronic countermeasures and airborne communication equipment. In order to operate over a wide band it is necessary that the characteristics of the antenna system be such that the radiation pattern is not appreciably changed over the frequency range involved. It is also necessary that the standing wave ratio be held to a minimum in order to prevent excessive losses.

The impedance of resonant wire antennas will vary with frequency, thus they will function adequately over only a narrow frequency band. As previously stated, in connection with the sleeve antenna, it is possible to maintain a uniform impedance over a wider frequency band by increasing the diameter of the antenna elements. This increase in diameter reduces the *Q* of the resonant system and this in turn causes a reduction in the reactive energy stored in the E and H fields adjacent to the antenna.

A cone antenna is a good example of increasing the diameter in order to obtain wide-band operation. The bandwidth of this type antenna can be further increased by the use of compensating reactances. By the use of these series and shunt reactances, it is possible to increase the operating frequency range of the antenna as much as six times.

Figure 7–29 shows two cone-type high-frequency antennas for use with ECM equipment. The electrical connections are brought to a receptacle at the bottom of the antenna. The antenna flange is designed to be mounted flush with the skin of the aircraft. A plexiglass dome or blister is used to cover the antenna in order to afford protection and to reduce wind resistance. The antenna illustrated in (A) should be mounted vertically to receive vertically polarized waves and horizontally to receive horizontally polarized waves. The antenna in (B) is similarly constructed except that the conical pickup element is mounted at an angle of 45 degrees to the mounting flange. This permits reception of either horizontally or vertically polarized waves for a single mounting position.

Some antennas of this type have a high-pass filter incorporated in their assembly that attenuates all frequencies below their designed operating range. Signals below the operating range are filtered out and do not

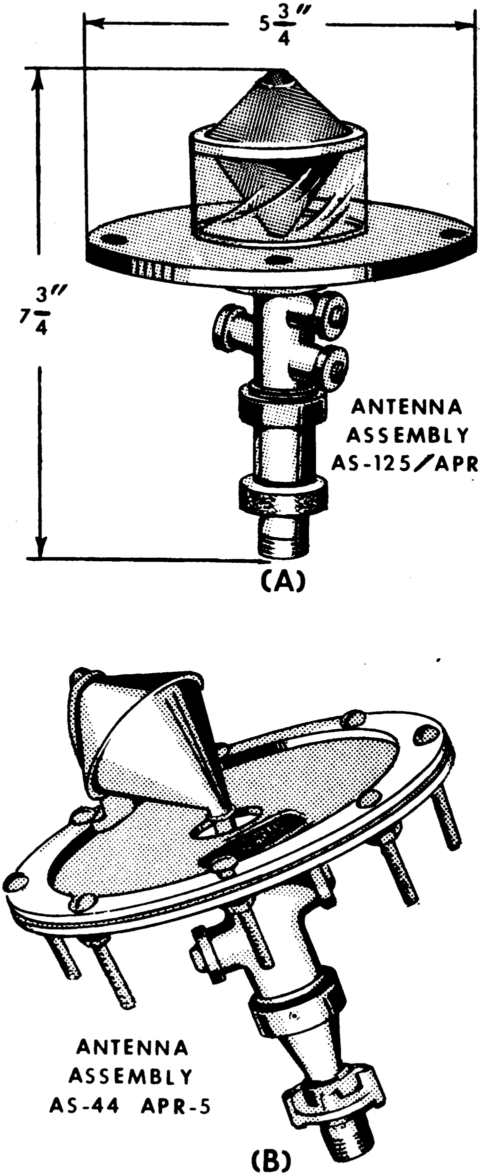


Figure 7–29.—Cone antennas.

reach the equipment. It is important when selecting antennas of this type, especially for ECM equipment, that a close check be made since an antenna that is equipped with this type filter will pick up a negligible amount of energy below its operating range.

The location of the antenna will vary depending upon the specific problems of the installation. It is extremely important to know that at all frequencies, especially UHF and microwave, it is possible to reduce the efficiency of an installation through the use of excess antenna cable lengths and connectors. Antenna runs should be kept to a minimum length. In most cases the *Handbook of Service Instructions* will specify maximum lengths. It is also important that either the designated type or a lower loss cable be used in all installation work.

A publication, *Index of RF Transmission Lines and Fittings,* NavShip 900–102B, previously mentioned in chapter 1, is available upon request. This publication contains charts of attenuation versus frequency for various types of transmission lines, assembly data for connectors, waveguide data, and other information. The charts will be helpful in determining attenuation losses for an antenna cable at a particular frequency. An example of attenuation loss is as follows: RG8/U cable has an attenuation of 9 db per 100 feet at 1,000 megacycles. If a 1,000-watt transmitter is connected to an antenna with 33⅓ feet of cable, the power at the antenna will be only 500 watts, thus, half the power will be dissipated by the cable. It is also possible to increase the losses of the system by using older type high-loss connectors. Even with the newer type low-loss connectors it is advisable to use as few as possible. This may be partially accomplished by using a minimum of right angle and through-the-bulkhead connectors; also, adapters should be used sparingly.

A modified cone antenna is shown in figure 7–30. This assembly consists of a cylindrical antenna with an inverted cone-shaped base fitted into a mounting flange. It

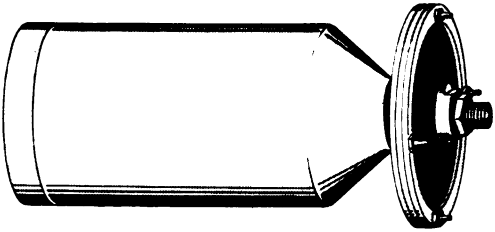


Figure 7–30.—Antenna assembly.

is designed as a jamming transmitter antenna and covers the frequency range of 200 to 550 megacycles.

A variation of the cone antenna is illustrated in figure 7–31. It is a flat, diamond-shaped metal plate enclosed in a streamlined plastic housing and is omnidirectional. The actual antenna is shown in (A) of the figure and the equivalent circuit of the antenna system is shown in (B).

An antenna of this type when operated at about 1,000 megacycles will provide efficient operation over a bandwidth of approximately 200 megacycles. Since this antenna is made of a flat piece of metal, it may be mounted in an airfoil type plastic housing that offers little wind resistance. This is especially desirable for high-speed aircraft.

The radiator is made diamond shaped, in order to obtain broad-band frequency characteristics, and is vertically polarized. Since it is end-fed, its impedance is comparatively high (140 ohms) and a matching section of transmission line is used to match this impedance to the 51-ohm transmission line. This section, which is a three-quarter-wave line, properly matches the impedances of the antenna and the antenna cable, and provides a low standing-wave ratio along the cable. The antenna requires no adjustment.

HORN ANTENNA.—An open-ended waveguide can be used to produce directional radiation. Some problems which occur are how to effectively match the waveguide to free space and how to achieve a narrow beam of radiation. Normally, the radiation pattern of an open-ended

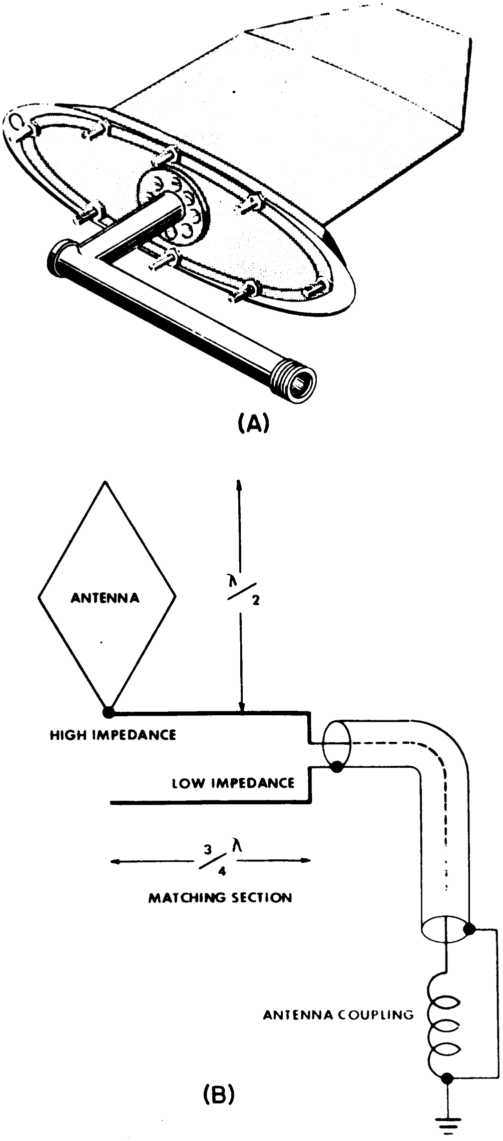


Figure 7–31.—UHF diamond antenna.

waveguide is broad in both the vertical and the horizontal planes. It is also an inefficient system because of the mismatch between the waveguide antenna and free space.

If the waveguide is flared out at a gradual angle as shown in figure 7–32 (A), a HORN antenna is obtained. The flare must be gradual so as to permit a better match between antenna and free space. At the same time, a sharper beam is obtained when a flared horn is used.

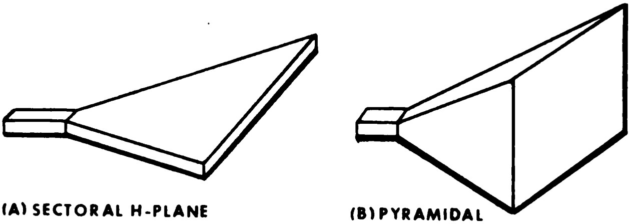
The horn is very practical at microwave frequencies since at these frequencies its physical size is not prohibitive. Since a resonant element is not involved, this type antenna is capable of wide-band operation. Horn antennas used in aircraft installations produce beam patterns that are not highly directive. In order to produce a highly directive beam the antenna would have to be longer than practical for aircraft. This disadvantage may be overcome by placing a LENS at the mouth of the horn.

For horn antennas as well as for dipole arrays and reflectors, a rule that may be kept in mind is that the larger the aperture of the waveguide or the larger the area of the array, the greater will be the gain and directivity. Since a large aperture is desired at the mouth, the length of the horn may become unwieldy if the gradual flare requirement is to be incorporated in the design.

A sectoral-type horn antenna is illustrated in figure 7–32 (A). Its field distribution is spherical in shape and depends upon the mode of operation of the waveguide feeding it and the flare angle of the horn. The spherical shape of the wavefront becomes flatter (elliptical) as the flare angle of the horn decreases. This horn has the greatest degree of directivity in the plane of the widest dimension.

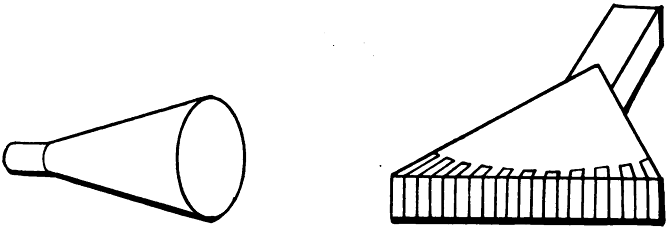
Two other types of horns to which the principles just described apply are shown in figure 7–32 (B) and (C). The pyramidal horn (B) has equal directivity in both the vertical and horizontal plane. The conical horn (C) produces a conical beam pattern.

As previously mentioned, horn antennas are often fitted



(A) SECTORAL H-PLANE

(B) PYRAMIDAL



(C) CONICAL

(D) LENS ANTENNA

Figure 7–32.—Horn antennas.

with a lens. These lens are made of plastic or conducting strips. When the lens is made of conducting strip, the strips are arranged parallel to the electric field and spaced slightly more than one-half wavelength apart. The wave entering the lens passes through the openings as though each opening is an individual waveguide section. (NOTE: The phase velocity of a wave traveling in a waveguide is greater than when traveling in free space.) A sectoral horn with a conducting-strip lens is shown in figure 7–32 (D). As you can see, the strips near the sides of the horn form longer sections of waveguides than those in the center. Thus, the wavefront will leave the horn in a plane of constant-phase since the phase velocity is progressively increased from the center of the horn toward the sides as the length of the individual waveguide sections increase.

The plastic-type lens, shown in figure 7–33, is affixed to a conical horn-type antenna. This particular antenna produces a beam width of 18 degrees at half-power points and has a gain of approximately 100. The lens is effective in reducing the sidelobe intensity by retarding the phase

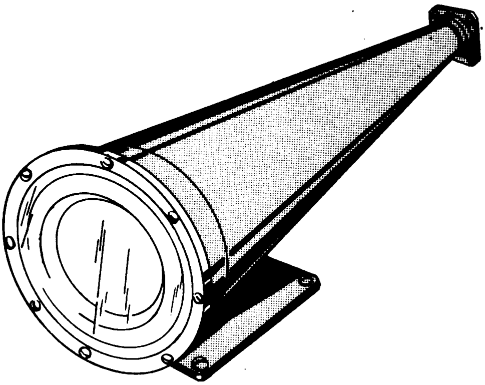


Figure 7–33.—Horn antenna with plastic-type lens.

of the energy leaving through the center of the lens. This causes the energy to leave the horn in a plane of constant-phase. It is interesting to note that while the two types of lens produce the same results they operate in somewhat opposite manner. The plastic lens retards the phase of the energy at the center and the conducting-strip lens accelerates the phase of the energy at the edge.

POLYROD ANTENNA.—Figure 7–34 shows a polyrod antenna that consists of a slender, tapered rod of special plastic. This rod is cemented to a piece of standard waveguide that it flared to match the square dimension of the large end of the rod. The principal mechanical features of this antenna are its streamlined aspect, its small size, and light weight.

It can be shown experimentally that very high frequency radio energy can be made to travel along a rod made of material having good dielectric properties in much the same way as light travels along a lucite rod. Some of the light traveling along the lucite rod is emitted from its longitudinal surfaces. Similarly, some of the radiofrequency energy traveling along the dielectric rod is radiated from its longitudinal surfaces. However, by appropriately tapering the rod, the amount of energy radiated per unit length may be made approximately

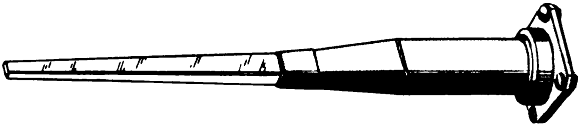


Figure 7–34.—A polyrod antenna.

equal. In doing so, an end-fire antenna is created which is roughly analogous to a linear array of dipoles spaced a quarter wavelength apart and fed at successive 90-degree intervals. For such an antenna array, the net energy radiated broadside to the rod will be a minimum. The rod, as seen from a point in a forward direction, will appear to be radiating a cigar-like beam of energy.

It is interesting to compare the characteristics of this antenna with those of the horn type shown in figure 7–33. While the half-power beam width of the horn is 18 degrees, that of the polyrod is 25 degrees. However, the spurious radiation from the polyrod is of greater magnitude than that from the horn. These factors cause the polyrod to be less efficient than the horn, but nevertheless, in some applications, it is more desirable because of its mechanical features.

QUIZ

1. A driven array consists of
   - a. two or more driven elements
   - b. one driven element and one parasitically driven element
   - c. only one driven element
   - d. two or more parasitically driven elements

2. A broadside array consists of two or more half-wave elements
   - a. mounted parallel to each other, spaced a quarter wave apart, and fed in phase
   - b. mounted parallel to each other, spaced a half wave apart, and fed 180 degrees out of phase
   - c. mounted parallel to each other, spaced a half wave apart and fed in phase
   - d. mounted along the same axis, spaced a half wave apart, and fed in phase

3. An end-fire array may consist of two or more half-wave elements
   - a. mounted parallel to each other, spaced a quarter wave apart, and fed in phase
   - b. mounted parallel to each other, spaced a half wave apart, and fed 180 degrees out of phase
   - c. mounted parallel to each other, spaced a half wave apart, and fed in phase
   - d. mounted along the same axis, spaced a half wave apart, and fed in phase

4. A collinear array is two or more half-wave elements
   - a. mounted parallel to each other, spaced a half wave apart, and fed 180 degrees out of phase
   - b. mounted parallel to each other, spaced a half wave apart, and fed in phase
   - c. mounted along the same axis and fed 180 degrees out of phase
   - d. mounted along the same axis and fed in phase

5. A parasitic array consists of
   - a. two or more driven elements
   - b. one driven element and one or more parasitically driven elements
   - c. only one element
   - d. all elements parasitically driven

6. For optimum operation of a parasitic array, the spacing should be
   a. 0.15 wavelength for the reflector and 0.1 wavelength for the director
   b. 0.1 wavelength for the reflector and 0.15 wavelength for the director
   c. 0.4 wavelength for the reflector and 0.5 wavelength for the director
   d. 0.5 wavelength for the reflector and 0.4 wavelength for the director

7. In a Yagi array, the
   a. director is about 5 percent shorter and the reflector about 5 percent longer than the driven element
   b. director is about 5 percent longer and the reflector about 5 percent shorter than the driven element
   c. reflector and director are about 5 percent longer than the driven element
   d. reflector and director are about 5 percent shorter than the driven element

8. The angle formed by the one-half power points of an antenna lobe and the transmitting station is known as the
   a. radiation angle
   b. lobe angle
   c. beam angle
   d. antenna angle

9. The input impedance of a folded dipole (not a part of a parasitic array) is
   a. 36 ohms
   b. 72 ohms
   c. 144 ohms
   d. 288 ohms

10. Increasing the diameter of an antenna will
   a. decrease its frequency range
   b. increase its frequency range
   c. increase radiation resistance
   d. decrease its center impedance

11. A ground plane is used with a vertical quarter-wave antenna when it is mounted several wavelengths above the ground to
   a. decrease the angle of radiation
   b. increase the angle of radiation
   c. decrease the antenna frequency range
   d. increase the antenna frequency range

12. The length of a ground plane should be
    a. one-third wavelength or less
    b. one-eighth wavelength or less
    c. one-quarter wavelength or greater
    d. exactly one-eighth wavelength

13. A slot antenna, mounted horizontally, will produce a wavefront which is
    a. horizontally polarized
    b. vertically polarized
    c. circularly polarized
    d. not polarized

14. The center input impedance of an omnidirectional half-wave slot antenna will be approximately
    a. 72 ohms
    b. 50 ohms
    c. 530 ohms
    d. 1,000 ohms

15. The range of a half-wave skin antenna is approximately the same as a
    a. half-wave fixed-wire antenna
    b. quarter-wave fixed-wire antenna
    c. three-quarter-wave fixed-wire antenna
    d. full wave fixed-wire antenna

16. The advantage of a cone antenna over a thin wire antenna is its
    a. greater gain
    b. greater frequency range
    c. higher *Q*
    d. greater directivity

17. An end-fed diamond-shape antenna will have an input impedance of approximately
    a. 70 ohms
    b. 560 ohms
    c. 140 ohms
    d. 51 ohms

18. A diamond-shape antenna, operating at 1,000 mc., will have an approximate bandwidth of
    a. 200 mc.
    b. 50 mc.
    c. 400 mc.
    d. 100 mc.

19. The polyrod antenna can be compared to a
    a. Marconi antenna
    b. lucite rod
    c. half-wave center-fed antenna
    d. half-wave end-fed antenna

20. The 10-cm. band would include which of the following frequencies?
    a. 9,375 mc.
    b. 2,880 mc.
    c. 15,950 mc.
    d. 300 mc.

21. To obtain the proper directivity in a parabolic reflector, it should have a diameter of
    a. a half wavelength
    b. a three-quarter wavelength
    c. many wavelengths
    d. a quarter wavelength

22. Two advantages of the cosecant-squared reflectors are a/an
    a. wide vertical beam width and wide horizontal beam width
    b. narrow vertical beam width and narrow horizontal beam width
    c. wide vertical beam width and equal power to all targets
    d. wide vertical beam width and nearly equal target returns

23. A longer flare on a horn antenna will give
    a. decreased directivity and increased gain
    b. increased directivity and decreased gain
    c. increased directivity and increased gain
    d. decreased directivity and decreased gain

24. In a horn antenna, the plastic lens reduces sidelobe power by
    a. retarding the phase of the energy leaving through the center of the lens
    b. retarding the phase of the energy leaving through the edge of the lens
    c. accelerating the energy leaving through the center of the lens
    d. stopping all energy leaving through the edge of the lens

CHAPTER 8

SERVOMECHANISMS

In the operation of electronic equipment it is often necessary to have the angular motion of a shaft follow accurately the motion of another shaft some distance away. If the shafts are not too far apart, the motion can be transmitted by means of a mechanical linkage. However, when the shafts are not close together, this method is, in general, cumbersome and uneconomical, particularly where large amounts of torque and power are to be transferred. A more flexible and more commonly used method is to convert the motion of one shaft into electrical currents, transfer these currents via wires, and convert them back to mechanical motion of a second shaft.

Two major applications of such conversions of shaft motion to electrical currents and back to shaft motion are remote indicator systems and servo systems. The term REMOTE INDICATOR SYSTEM refers to a device used for transmitting data or information between two points. Examples are the reading of a meter dial, the position of an antenna in azimuth or elevation, temperature readings, and similar information. For such indications very little torque and energy are involved. These systems are included under the general term SYNCHRO SYSTEMS and will not receive extensive coverage in this chapter since they are covered in the Navy Training Course, *Basic Electricity*, NavPers 10086, chapter 17.

The term SERVO SYSTEM refers to a mechanism whereby a controlled shaft usually delivers much more power than the controlling shaft, and the action of the power-amplifying device is regulated by the ERROR or discrepancy in angular position between the two shafts.

The term SERVOMECHANISM refers to a large variety of power-amplifying devices. It is frequently necessary to control the position of a device in accordance with some function of a signal supplied by a controlling instrument. If the power required to operate the device is large compared with the power available from the controlling instrument, power-amplifying means must be provided. When the amplification of the controlling signals is carried out by a machine or automatic device rather than by a human operator, the complete system is known as an AUTOMATIC CONTROL SYSTEM, and the element that amplifies the controlling signals and operates the device is known as an AUTOMATIC CONTROLLER. The servo controller is the device in which the input is the error and the output controls the servomotor. This device is also known as a servo amplifier.

Automatic control systems can be divided into two types, depending upon the source of the signals that actuate the controller. If the signals supplied to the controller are determined not only by the controlling instrument but also by the position of the device being controlled, the system is known as a CLOSED-CYCLE control system. This system is the one most frequently used since it is seldom possible to find a power-controlled mechanism in which the cycle is not closed either mechanically, electrically, or through a human link. If the signals supplied to the controller are not a function of the position of the device being controlled, the system is known as an OPEN-CYCLE control system. An example of the open-cycle control system is the electric type fuel gage on an automobile. The float in the fuel tank moves the arm of a potentiometer and causes the arm to pick off a voltage that is proportional to the height of the fuel in the tank. The gage is a simple voltmeter that indicates the amplitude of the voltage, thus the potentiometer movement is followed by a movement of the gage indicating needle.

The term servomechanism is used to denote a certain type of closed-cycle automatic control system. If the controlling signals to the system change frequently and rapidly and the primary problem is to cause the controlled device to follow these signals, the system is called a SERVOMECHANISM. If the controlling signals do not change, or change relatively slowly, and the primary problem is to prevent extraneous causes from affecting the controlled element, the system is called a REGULATOR. Servomechanisms are used to steer ships, to control airplanes, to automatically tune radio receivers, to position guns, in numerous telemetering and computing applications, and, in general, in those automatic control applications where accuracy of reproduction is of primary importance.

SERVOMECHANISM TERMS

The following terms have been applied to servomechanisms. The quantities defined are well established and, although they have not been standardized, are understood by those active in the field.

INPUT.—The input is the controlling quantity. It may be the displacement of a shaft, a voltage or current, a temperature level, and so forth.

OUTPUT.—The servo output is the position or state of the controlled quantity. It may be any of the quantities listed in the definition of the INPUT.

ERROR.—The servo error is the difference between the servomechanism input and the output.

ERROR DETECTOR.—The error detector is the device which compares the input with the servomechanism output.

SERVO CONTROLLER.—The servo controller is the device in which the input is the error and the output controls the servomotor.

SERVOMOTOR.—The servomotor is the prime mover of the servomechanism. It is actuated by the servo controller and controls the servo load or output.

SERVO LOAD.—The servo load is the quantity upon which the servomotor operates.

COMPONENTS OF AN ELECTROMECHANICAL SERVOMECHANISM

The essential components of a servomechanism are the data transmission system, the servo controller, and the servomotor. These components are shown in the block diagram of figure 8–1. The function of each component will be discussed. While the servo load, strictly speaking, is not a component of the servomechanism, its characteristics have an important bearing on the design and operation of the servo system.

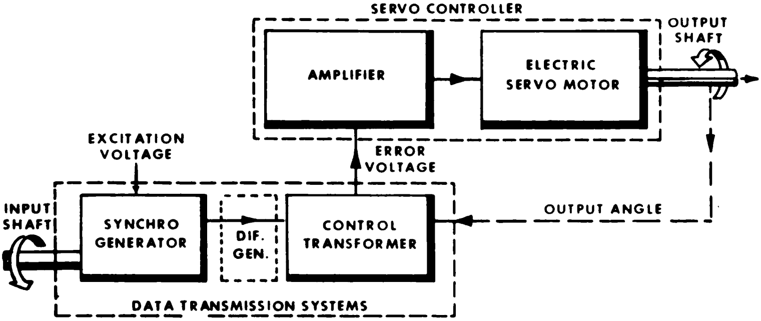


Figure 8–1.—Simplified block diagram of a servomechanism.

DATA TRANSMISSION SYSTEMS.—In some servomechanisms the data transmission system may consist solely of an error detector (control transformer). In many servo systems, both the physical position of the servo output and the origin of the signals comprising the servo input may be remote from one another and from the servo controller. The function of the data transmission system is to transmit a signal from the input signal generator to the control transformer. This signal is combined with the servo output in the error detector (control transformer) which determines the magnitude and direction of error. This error signal is then applied to the servo controller.

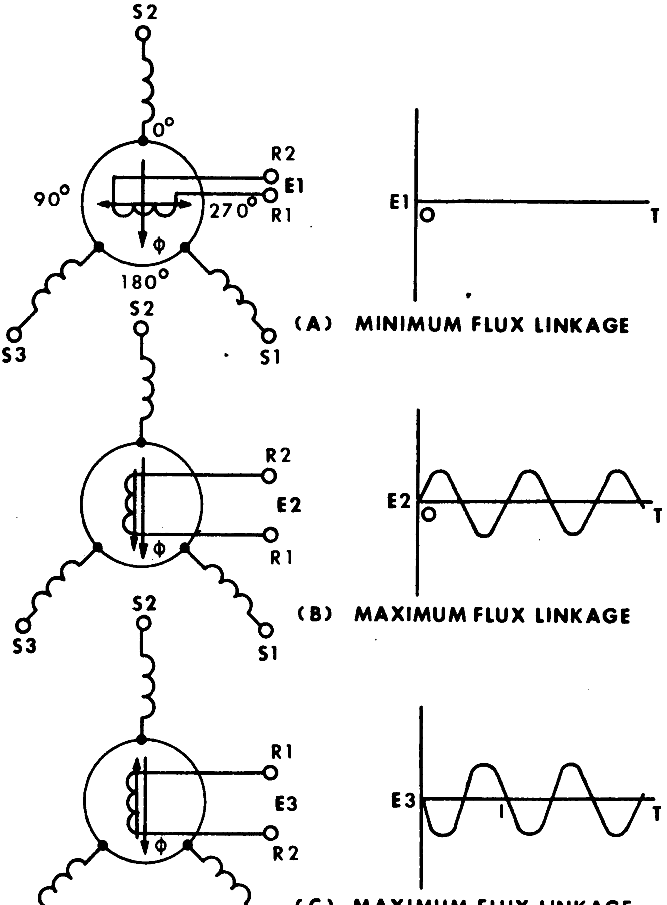
One component of the data transmission system is the error detector which is the mechanism for comparing the input and output functions. The error detector may be resistive (potentiometer), capacitive, or one of several types of magnetic devices.

A commonly used magnetic error detector is the synchro. The synchro has been developed to the point that it is characterized by relatively high accuracy, low noise level, reasonably small driving torques, and excellent life. A primary advantage over other types of error detectors is its unlimited rotation angle; that is, both the input and the output to the synchro may rotate through unlimited angles. Among the disadvantages of synchros are the large size necessary to maintain high accuracy, the high power consumed, and the fact that the output supplied to the servo controller is always an alternating voltage modulated by the servo error angle and must be demodulated in either the servo controller or the servomotor.

The synchro data transmission system is comprised of a synchro generator, a synchro receiver, and in some cases a differential generator is included. (See *Basic Electricity,* NavPers 10086, chapter 17, for the theory of operation and function of a differential generator.) The synchro generator transforms the motion of its shaft into electrical signals suitable for transmission to the synchro receiver which comprises the error detector.

The stator of the generator consists of three coils spaced 120 electrical degrees apart. The voltage induced in the stator windings is a function of the position of the salient pole generator rotor. These voltages are applied to the three similar stator windings of the synchro receiver or control transformer. The voltage induced in the rotor of the synchro receiver depends upon the relative position of this rotor with respect to the direction of the stator flux. The variation of the synchro-receiver output voltage as a function of the rotor position relative to an assumed stator flux direction is shown in figure 8–2. While there are two positions of the rotor, 180 degrees apart, where the output voltage is zero, only one corresponds to a stable operating position of the servo.

Another type of error-detecting element commonly used



(A) MINIMUM FLUX LINKAGE

(B) MAXIMUM FLUX LINKAGE

(C) MAXIMUM FLUX LINKAGE. ROTOR DISPLACED 180° FROM (B) ABOVE

Figure 8–2.—Induced voltage in synchro control transformer rotor.

in the data transmission system is the potentiometer. Potentiometers are generally used only where the input and the output of a servomechanism have limited motion; they are characterized by high accuracy, small size, and the fact that a d-c or a-c voltage may be obtained as the output. Their disadvantages include limited motion, a life problem resulting from the wear of the brush on the potentiometer wire, and the fact that the voltage out of the potentiometer changes in separate steps as the brush moves from wire to wire. A further disadvantage of some potentiometers is the high driving torque required.

Capacity devices have not been widely used to date as error detecting devices in servomechanisms.

MULTIPLE-SPEED DATA-TRANSMISSION SYSTEMS.—The static accuracy of a servomechanism is frequently limited only by the accuracy of the data-transmission system. The accuracy of the data-transmission system may be increased considerably by employing a multiple-speed data-transmission system along with the single-speed system. The error-detector elements of the multiple-speed transmission system rotate at some multiple of the shaft being controlled, while the elements of the single-speed transmission system operate one to one with respect to the controlled shaft.

The schematic diagram of a multiple- and single-speed system is shown in figure 8–3. If the input shaft turns through 1 degree in this example, the single-speed synchro transmitter also is rotated 1 degree while the multiple-speed unit is rotated 10 degrees. The synchro receivers associated with each of these transmitters are geared in similar ratios with respect to the servo output shaft. A 1-degree error between the position of the input and output shafts therefore produces a relative rotor displacement of 1 degree in the single-speed synchros and 10 degrees in the multiple-speed synchros. If the relation between the rotor displacement and output voltage is linear, the error signal from the multiple-speed system

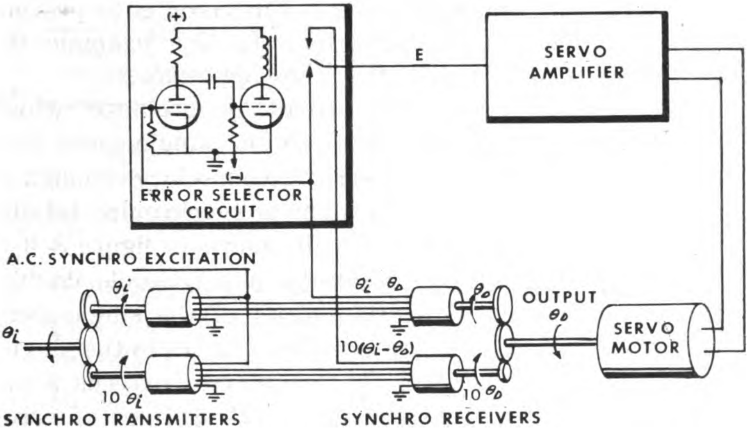


Figure 8–3.—Multiple-speed synchro data-transmission system.

is ten times that from the single-speed system. This amplification of the error signal in the data-transmission link reduces the signal amplification required in the servo controller. If the synchro has an inherent error of 0.1 degree with respect to its own shaft, the consequent servo error introduced by a single-speed data-transmission system will be of corresponding magnitude, but the consequent servo error introduced by a 10-speed data-transmission system will be only one tenth as great or 0.01 degree.

A disadvantage of using a multiple-speed error detector lies in the possibility of the system falling out of step and synchronizing in a position differing from the correct position by an integral number of revolutions of the multiple-speed synchro. In the example shown in figure 8–3, if the output shaft were held fixed and the input shaft rotated 36 degrees, the 10-speed synchro transmitter would turn through one complete revolution. At this point the error-signal from the multiple-speed error detector would be zero. If the output shaft were then released, the system would operate in stable fashion with a 36-degree error between the input and output shafts. The purpose of using a single-speed error detector along with the multiple-speed detector is to prevent this ambiguous synchronization. In this example the ambiguity is prevented in the following manner:

An error-signal-selector circuit is provided which switches control of the servo to the single-speed data system whenever the servo error becomes large enough to permit the multiple-speed system to synchronize falsely. The type of error-selector circuit shown in figure 8–3 is frequently used. It is essentially a single-pole double-throw relay actuated by the output of the single-speed error detector. This relay switches control to the single-speed synchros whenever the servo error exceeds about 12 degrees, which corresponds to 120 degrees of error displacement of the multiple-speed synchro rotors.

With the error-signal-selector circuit shown, it is still possible to have a single ambiguous synchronizing point in the system, because the multiple-speed synchros operate at an even multiple of the single-speed synchros. This ambiguous synchronization can occur when the servo input and output shafts are 180 degrees out of phase. In this position both the single- and multiple-speed synchro systems are at null voltage positions and control of the system is retained by the multiple-speed units. One way of eliminating this false synchronization position is by operating the multiple-speed synchros at an odd multiple of the single-speed synchros. While the 180-degree null position of the single-speed synchros still coincides with a null voltage position of the multiple-speed synchros, the corresponding servo position is an unstable one and the servo will not remain at this point.

The disadvantage of using multiple-speed error detectors is the need of an additional synchro system and switching circuit if false synchronization is to be avoided. The increased servo accuracy resulting from the amplification of the error signal and the effective reduction of inherent synchro errors, however, account for the wide use of these multiple-speed data systems.

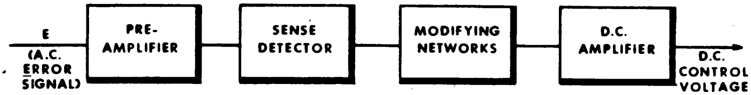


Figure 8–4.—Block diagram of servo amplifier.

SERVO AMPLIFIER.—The functions of the servo amplifier are:

1. To supply required amplification of the error signal.
2. To transfer the error signal into suitable form for controlling the servomotor.
3. To provide special characteristics that may be necessary to obtain stable, fast, and accurate servo operations.

Every servo amplifier provides one of these functions and many provide all three. A block diagram of a servo amplifier which performs all three of these is shown in figure 8–4. The input to this amplifier is represented as an a-c signal, the magnitude of which is proportional to the magnitude of the error between the input and the output and the phase of which is a function of the sign of the error. The first portion of the servo amplifier is an a-c amplifier that supplies the gain necessary for proper operation of the system. The second component of this amplifier is a sense-detecting rectifier or demodulator. This element transforms the a-c error signal into a d-c signal, the magnitude of which is proportional to the magnitude of the servo error and the polarity of which is determined by the direction of the servo error. The final component of the amplifier is a d-c amplifier that raises the d-c signal to the power level necessary to control the servomotor.

In many control systems the servo amplifier is not required to perform all of the functions indicated in the amplifier shown in figure 8–4. For example, if the servomotor is to be controlled by an a-c voltage of the same frequency as the a-c error signal, the detector and rectifier might be eliminated and the d-c modifying network replaced by a-c networks or by devices that produce the desired effect in other ways.

Thus the functions of the servo amplifier and its exact form may vary widely in different control applications. Examples of actual amplifiers are given in other sections of this chapter.

SERVOMOTORS.—The function of the servomotor is to supply the torque, power, and dynamic characteristics required to position the servo load satisfactorily. The servomotor itself is controlled by the servo amplifier. Ideally, a servomotor should require only small amounts of power from the amplifier, should be capable of accelerating rapidly, and should be characterized by small size and weight, long life, small time lags, and an adequate speed range. Some of the electric motor types frequently used in servomechanisms are indicated as follows:

The 2-phase induction motor is the most widely used a-c servomotor. One phase of this motor is excited from a constant reference source; the other phase is excited in quadrature (at right angles) with this reference phase by the servo amplifier output. The motor torque is proportional to the amplitude of the control voltage and its direction depends on the relative phase of the excitations on the control and reference phase.

The most frequently used d-c servomotor is the shunt motor. It is controlled by varying either the armature or field current. The uncontrolled current is usually maintained constant to preserve a linear relationship between the motor output torque and the voltage or current input. The field windings are usually two differentially wound coils to facilitate direction control of the field current by the servo amplifier. The field current is usually controlled with receiving type vacuum tubes, while the larger armature currents usually require thyratrons or generators as current regulators. In some very small motors the field may be supplied by a permanent magnet.

Certain types of shaded-pole induction motors have been developed which are also suitable as servomotors, but which are not widely used at present. In each of these motors the main field winding usually is excited from a constant a-c source and the motor output-torque amplitude and direction are controlled by an alternating voltage and direct voltage applied to a shading winding. The alternating control voltage is either in phase or 180 degrees out of phase with the main field voltage. As in the case of the 2-phase induction motor, the sense detector may be unnecessary if a synchro error detector is used in the servo system.

Universal motors and clutch mechanisms are sometimes used as servomotors, but in general they have not been applied as extensively in control systems as the types of motors mentioned.

CIRCUIT FUNCTIONS AND ANALYSIS OF SERVO AMPLIFIERS

Phase Discriminators

A phase discriminator or, as it is sometimes known, a phase sensitive rectifier, will mix two a-c voltages and provide a d-c control voltage or current of reversible polarity at the output. The two voltages to be mixed will in most applications be in phase or 180 degrees out of phase.

DIODE PHASE DISCRIMINATOR.—A typical diode phase discriminator is shown in figure 8–5. As illustrated, an a-c excitation voltage serves as the reference voltage to the discriminator. This voltage must come from the source supply that is supplying a-c excitation to the synchro system that will create the voltage to be compared to the reference. The plates of the two diodes are supplied with this reference voltage in such a manner that the two plates will be in phase. Assuming that there is no error signal from *T*2 into the plates of the diodes at the time the plates are on a positive half cycle the two diodes will conduct equally. The voltage produced across *R*1 and *R*2 are at equal potential in respect to ground. The a-c component is removed by capacitors *C*1 and *C*2.

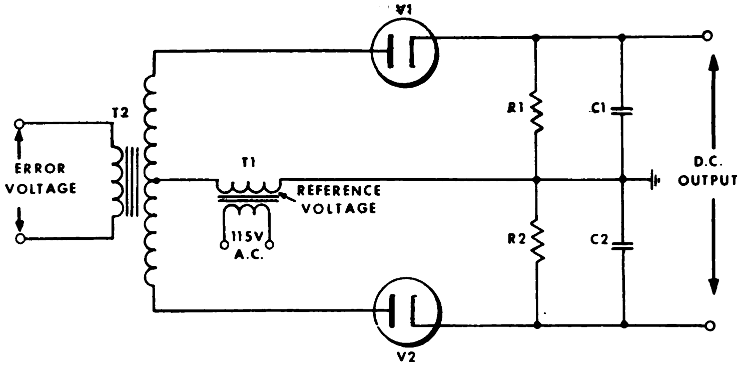


Figure 8–5.—Phase sensitive rectifier using diodes.

The output is zero as long as no signal is introduced by *T*2.

For a further understanding of the circuit, assume that a control transformer is controlling the circuit and is manually turned to a position to produce an a-c voltage of a definite phase relationship to the reference and of an amplitude determined by the amount of rotor change. If the phase relationships of the two voltages are such that the plate of *V*1 has a positive half cycle of error signal at the same instant that the reference voltage on the plates is on its positive half cycle, *V*1 will conduct more than *V*2. At this same instant, the plate of *V*2 will have a negative half cycle of error signal which will reduce its conduction. The cathode of *V*1 is now positive in respect to the cathode of *V*2, due to the charges on *C*1 and *C*2. The change in amplitude in this direction will actuate a control circuit which will move the output shaft in the proper direction. The correction speed of the output shaft will be relative to the magnitude of the d-c output voltage. When in the proper position, the output shaft will, through a feedback device, cause the error voltage to drop to zero, at which time the output voltage will return to its static level, stopping the controlled action.

If the control transformer rotor were turned in the opposite direction, the phase of the voltage applied to the primary of *T*2 would be changed by 180 degrees. At this time the plate of *V*1 would now be on a negative half cycle of error voltage at the same instant the two plates would be on a positive half cycle of the reference voltage. Thus, the conduction of *V*1 will be reduced and the conduction of *V*2 will increase above its static level. The d-c output will change so that the cathode of *V*2 will be positive in respect to the cathode of *V*1. This new polarity of d-c voltage will cause the controlled action to be in a direction opposite to that previously given. This causes the output shaft to move in the opposite direction.

TRIODE PHASE DISCRIMINATOR.—A phase discriminator using triode tubes is shown in figure 8–6.

In this circuit, as in the previous one, the plates of the tubes are supplied with the a-c reference voltage in such a manner that the plates will be in phase. For purpose of explanation assume that no error signal is present at *T*2. When the plates of *V*1 and *V*2 are positive the two tubes will conduct equally. The plate current that flows will set up fields in the d-c motor exciter windings that will be equal and opposite, therefore, the fields will cancel and

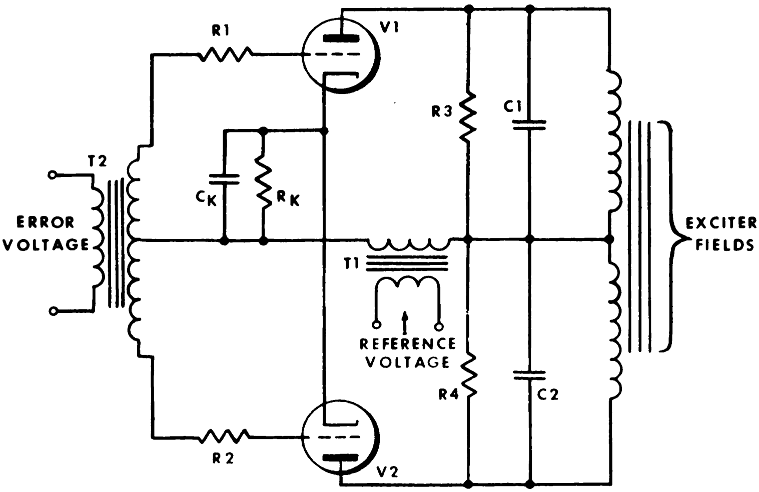


Figure 8–6.—Servo amplifier.

produce no output. When the plate voltages are on a negative half cycle, *C*1 and *C*2 will be discharging through their respective exciter windings to maintain a constant d-c current through the windings.

If an error signal is introduced into the primary of *T*2 of such a phase relationship that the grid of *V*1 will be positive at the same instant that the plate of *V*1 is positive, the following conditions will exist:

1. On this half cycle the conduction of *V*1 will be increased above its static value.
2. The heavier plate current will cause a stronger field to be created in the upper exciter winding.
3. At this same instant, since the grid of *V*2 is on a negative half cycle, its average conduction will be reduced to a level below that of its static value.
4. The lower level of plate current will cause a weaker field to be produced in the lower exciter winding.
5. Since the magnetic fields produced in the exciter windings are no longer of equal amplitudes, they can no longer cancel each other.
6. The exciter will produce an output voltage of a polarity controlled by the polarity of the resultant field and of an amplitude controlled by the relative strength of this resultant field.
7. The exciter output will cause the proper mechanical actions necessary to reduce the amplitude of the error to zero.
8. As the error signal is reduced to zero, the current conduction through *V*1 and *V*2 will again be balanced and the exciter fields will be equal and opposite and cancel each other, reducing the exciter output to zero and stopping the mechanical action. Resistors *R*1 and *R*2 prevent excessive grid current when the error angle is large.

FULL WAVE DIODE BRIDGE DISCRIMINATOR.—A tilt control circuit using a full wave diode bridge discriminator

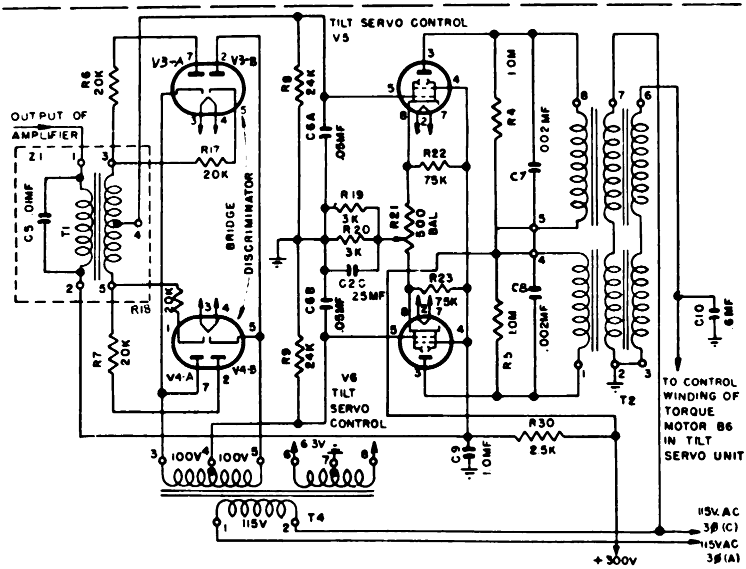


Figure 8–7.—Discriminator and tilt-servo control circuit.

is illustrated in figure 8–7. This discriminator consists of tubes *V*3 and *V*4 and associated circuits.

This circuit will produce a d-c voltage with its polarity controlled by the phase relationship of the input error signal with respect to the reference voltage. The amplitude of the d-c output control voltage will be determined by the amplitude of the error signal.

Assuming no error signal is present the circuit will perform as follows:

At the instant when the voltage at terminal 3 of *T*4 is negative and terminal 5 is positive, current will flow from terminal 3 to the cathode of *V*3–*A* and then to its plate. From the plate of *V*3–*A* it flows through *R*6 to terminal 3 of *T*1 and through *R*17 to the cathode of *V*3–*B*, through the tube and back to terminal 5 of *T*4. On the alternate half cycle when terminal 3 of *T*4 is positive and terminal 5 is negative, current will flow from terminal 5 of *T*4 to the cathode of *V*4–*B*, through the diode to

*R*7 and to terminal 5 of *T*1, from terminal 5 of *T*1, through *R*18 to the cathode of *V*4–A, through *V*4–*A* back to terminal 3 of *T*4. Note that during this entire cycle no current will flow through *R*8 and *R*9, since currents tend to flow in both directions AT THE SAME TIME. They are equal and opposite; thus, they cancel and the voltage across the output terminals will be zero.

An error signal introduced into *T*1 will be of a definite phase relationship with respect to the reference voltage; it will be either in phase or 180 degrees out of phase. If it is assumed that terminal 3 of both *T*1 and *T*4 are in phase and on a positive half cycle, current will flow as follows: From terminal 5 of *T*1 through *R*18 and *V*4–*A* (cathode to plate), to terminal 3 of *T*4 then to the center tap terminal 4 of *T*4. From terminal 4 of *T*4 through *R*9 and *R*8 to the center tap terminal 4 of *T*1 and back to terminal 5 of *T*1. On the alternate half cycle, terminal 3 of both transformers will be negative. Current will then flow from terminal 3 of *T*1 through *R*17 and *V*3–*B* (cathode to plate), to terminal 5 of *T*4 to the center tap terminal 4. From terminal 4 of *T*4 through *R*8 and *R*9, on to terminal 4 of *T*1 and back to terminal 3 of *T*1. For a complete cycle current has passed through *R*9 and *R*8 as a d-c current, causing the bottom end of *R*9 to be negative in respect to the top of *R*8. *C*7 and *C*8 will remove any a-c component of voltage which may appear at the output. If terminal 3 of *T*1 and *T*4 had been out of phase, the direction of current flow through *R*9 and *R*8 would have been in the opposite direction, so that the output polarity of voltage would have been reversed.

Induction-Motor Type Servo System

A 2-phase induction motor is frequently used with servo systems in aviation installations. Figure 8–8 illustrates by way of a block diagram a basic circuit using this type motor.

The action that takes place is as follows: An alternating voltage of fixed amplitude is applied to coil *A*.

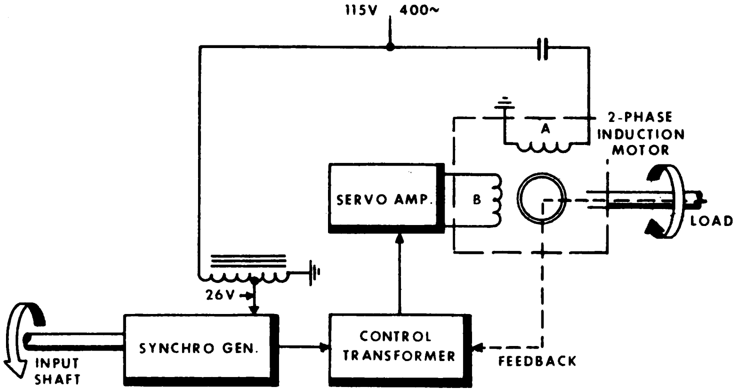


Figure 8–8.—Induction-motor type servo system.

Capacitor *C* is connected in series with coil *A* for the purpose of providing a phase shift so that the voltage applied to this coil is nearly 90 degrees out of phase with the synchro voltages. Thus, the output of the servo amplifier, which is connected to coil *B*, will be 90 degrees out of phase with coil *A*. This action provides the necessary phase difference between the coils for proper operation of the motor.

When the input and output shafts are displaced from each other an error voltage will be produced. As a result of this error voltage, the servo amplifier will supply a current to coil *B*. This current in conjunction with coil *A* produces a rotating flux that drives the induction motor in the direction to reduce the error to zero.

When the error voltage is zero only coil *A* will produce a magnetic flux in the induction motor. Since this alternating flux does not rotate, no torque is produced by the motor.

The servo amplifier in this circuit functions as an audio amplifier. It provides an amplified output of the error voltage that is in phase with the error voltage.

Antihunt Devices

Servomechanisms and associated amplifiers will necessarily have high-power gain in order to provide fast response from input to output shaft. Servomechanisms must be capable of supplying sufficient power to create high enough torque to overcome load inertia and insure accurate controlled motion. A servo that does this causes the output shaft to approach a new position at a high rate of speed. When the output shaft reaches the correct position at full speed, mechanical inertia will cause it to run beyond the correct point, commonly called OVERSHOOT. This overshoot creates an error signal of opposite phase at the servo input which causes the motor rotation to reverse the output shaft's direction. The results of this action are:

1. The reversed motion drives the output shaft back through its correct position.
2. Motor rotation will again reverse to the original direction.
3. The amplitude of output shaft motions will gradually decrease each time it passes through its correct position.

This oscillation is usually termed HUNTING. It must be prevented so as not to cause damage to the output shaft's mechanical components. Also, since the output shaft may take a relatively long time to stop or dampen out these oscillations, this hunting will with all probability result in very poor servo operation.

MECHANICAL FRICTION METHOD.—One method of creating antihunt action is to create a high mechanical friction on the output shaft, but this system has many disadvantages. They are:

1. Much of the servomechanism output power is lost in frictional heat, thus creating a very inefficient system.
2. Such a system requires much heavier and larger mechanical components.
3. The positioning of the output shaft is less accurate.
4. The degree of error between input and output shaft is quite large.

5. The speed of the output shaft is slower than that desired of an efficient servo system.

INERTIA DAMPERS.—Inertia dampers such as are used on synchro receivers are practical only on the very smallest of servo systems. This is true because the inertia of the damper must be many times that of the output shaft and its mechanical load.

LIMITER TYPE ANTIHUNT DEVICE.—A simple electrical means of antihunt is to employ a limiter in the error amplifier ahead of the discriminator. A very low plate supply voltage applied to one of the error voltage amplifier stages will limit the gain of the stage to a very low level. In such a system a low amplitude error signal produces an amplified output. This output gives a definite speed of output shaft motion which is comparatively low. Increasing the error signal amplitude causes no further increase in output because of the limiting action of this amplifier stage. The motor and output shaft speed will thus have no increase. The output shaft speed will necessarily be quite slow and the mechanical load will be great enough to prevent overshoot. The main disadvantage of this system is: It causes very slow response to an error angle between input and output shafts due to the low servo gain. This form of antihunt has been used with azimuth stabilization systems where output speed need not be fast.

HIGH-SPEED ANTIHUNT DEVICE.—Figure 8–9 illustrates the AN/APS–31/33 tilt-stabilization system which contains a high-speed antihunt device.

An a-c generator is geared mechanically to the servo torque motor and has an output voltage when the torque motor turns. This is the damping signal generator shown in figure 8–9. The a-c voltage out of this generator is adjusted to a value equal to the error signal produced with only a few degrees of angular displacement between input and output shafts. This ANTIHUNT or DAMPING voltage is fed to the servo amplifier in such a manner as to be out of phase with the error voltage at all times. As the servo torque motor turns the output shaft to its new position, the amplitude of error signal will be constantly decreasing to a null of voltage. At some point close to zero angular displacement between input and output shafts, the antihunt a-c voltage will be equal in ampli-

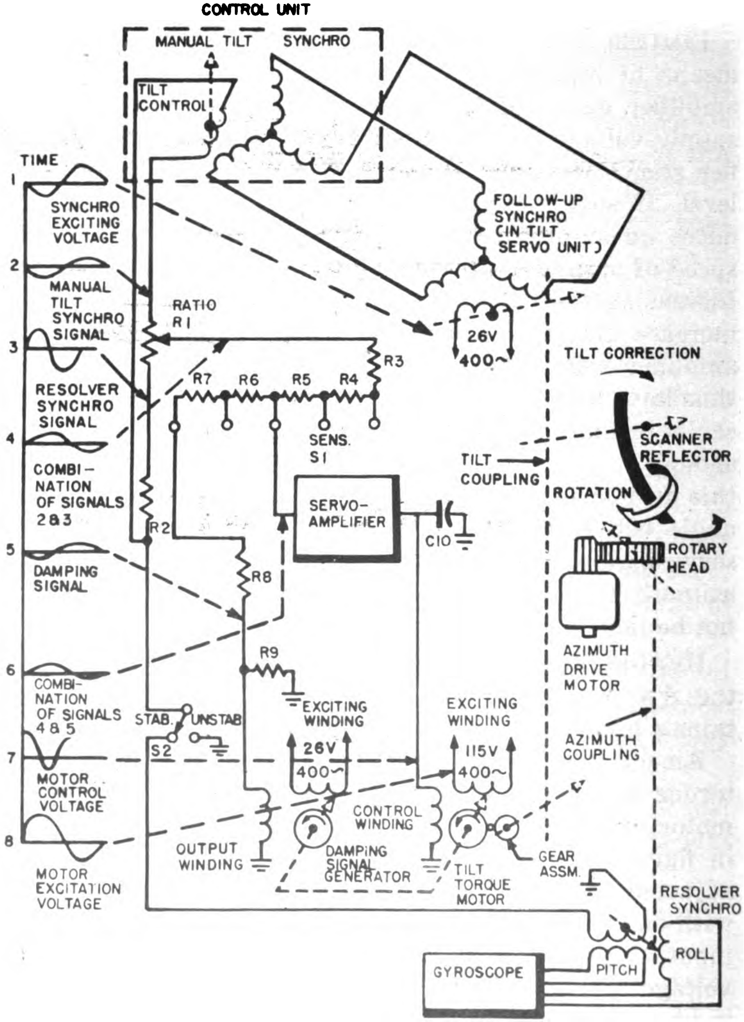


Figure 8–9.—Tilt stabilization with high-speed antihunt.

tude but out of phase with the error voltage, so that the signal into the servo amplifier is cancelled. The inertia of the output shaft mechanism carries the output shaft beyond this point to where the phase of the signal is reversed and starts to increase in amplitude due to the damping signal generator's output exceeding the error signal. This causes a reversal of the output voltage from the servo amplifier to the torque motor. This reversal of torque causes a braking action which stops the torque motor and the load. This also stops the damping signal generator and no further error is applied to the amplifier. Some provision must be made to control the exact amplitude of the input signal to the servo amplifier so that the gain of the amplifier may be controlled. This may be done by varying the antihunt or damping generator output so as to cancel out the error signal at the proper angular displacement.

If the antihunt or damping signal is LOW in amplitude as compared to the error signal at the correct cancellation angle, the power to the torque motor will not be removed at the proper angular displacement point. This will cause the output shaft to overshoot the proper angle and oscillations or hunting will occur. These oscillations will gradually diminish to the correct position of the output shaft. If the antihunt or damping signal is HIGH in amplitude as compared to the error signal at the proper cancellation angle, the error signal will be cancelled at too great an error angle, or too soon. The output shaft will be stopped with some degree of error angle remaining between input and output shafts. As the output shaft stops, the a-c damping voltage will fall to zero and the remaining error signal will cause the shaft to turn slowly to the correct point.

DERIVATIVE TAKING ANTIHUNT NETWORK.—A derivative taking network used as an antihunt device is shown in figure 8–10. The circuit shown is the first voltage amplifier in a servo amplifier used to drive a 2-phase

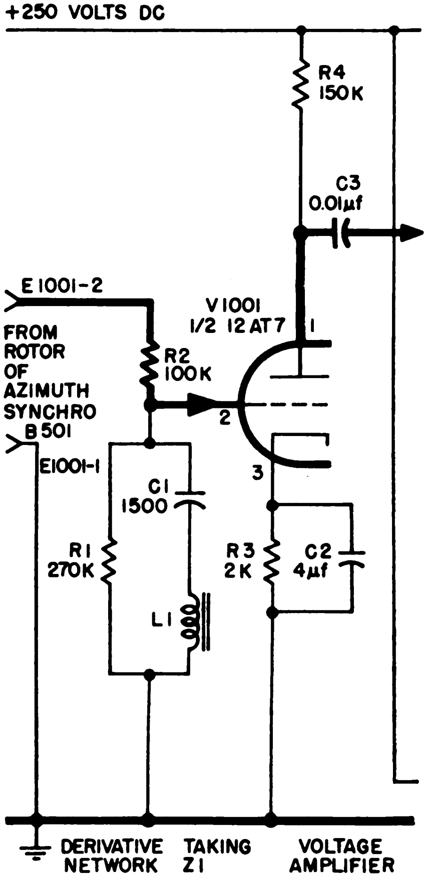


Figure 8–10.—Derivative taking network.

induction motor. The derivative network *Z*1 consists of *C*1 and *L*1.

This network is a high impedance series resonant circuit which suppresses hunting and overshoot caused by the inertia of the followup system. This is accomplished by introducing a phase lead in the servo system. During the reversals which are characteristic of sector-scan operation in a radar azimuth servo system, the torque required to overcome the lag in the system is quite high.

The rotor output of the synchro control transformer is fed through voltage dividing resistor *R*2 to the mesh composed of resistor *R*1 in parallel with capacitor *C*1 and inductor *L*1. The reactive elements form a series-resonant branch at the supply frequency and offer very low impedance. At any other frequency, they appear inductive or capacitive and cause the output of the derivative network to lead in phase, thus the load anticipates any change in input speed and direction.

Review of a Servomechanism System

As you review the operation of a servomechanism system use figure 8–1 as a reference. The connections and functions of the basic servomechanism components are as follows: Consider that when the input shaft is turned, an error voltage is produced. This voltage is taken from the rotor of the control transformer and applied to some type of servo amplifier. The amplifier will supply power to the servomotor which will turn the output shaft. The output shaft will operate the desired device and is also mechanically connected back to the control transformer.

The basic operating sequence is as follows: When the input shaft is moved, a pattern of voltages will be transmitted by the synchro generator to the control transformer stators. An error voltage will be supplied to the amplifier from the control transformer rotor. The phase of this voltage with respect to the reference voltage depends upon which direction the input shaft is turned. The amplitude depends upon the amount of displacement between the input and output shafts. The servo amplifier develops a control voltage or current of the proper phase and amplitude to be applied to the control winding or windings of the servo. The control circuit will cause the servomotor and control transformer rotor to turn in the same direction and the same number of degrees as the input shaft has been turned. Once the rotor position of the control transformer differs from the synchro generator rotor position by 90 degrees, the input to the amplifier is zero and the system returns to a static state.

It is important that you know how to make the proper adjustments to servomechanisms in order to obtain from them the proper performance. Adjustments must be made when new equipment is installed, when components such as control transformers, differential synchros, servo amplifiers, tubes, or other parts of the servo system are replaced, and during normal service as a result of variations in the system due to the usage of the parts. These adjustments will provide changes in gain, phase, and balance of the servo amplifier and also the alinement or zeroing of the synchros of the system. The zeroing of synchros in general has been covered in chapter 17, *Basic Electricty,* NavPers 10086, and will not be covered in this chapter. The specific instructions for zeroing the synchros of each specific type equipment will be found in the *Handbook of Service Instructions* for that equipment.

PHASE ADJUSTMENT.—After the servo system has been zeroed or alined, it is then necessary to ascertain if the phase relationships are correct. Phase shift adjustments are frequently provided in a servo system to compensate for any undesirable phase shift caused by the amplifier or the synchro data transmission system. When a system uses a 2-phase servomotor the controlled winding voltage must be 90 degrees out of phase with the fixed voltage winding. Many servo systems provide a means of adjusting the phase difference between these voltages. There are various methods of making this adjustment. Some equipments provide a means of adjusting the phase of the fixed voltage winding. This is accomplished by varying the capacitance of a capacitor that is connected in series with the fixed voltage winding. Other servo systems use an amplifier that contains a variable R–C or R–L network which, when adjusted, varies the phase of the controlled voltage winding.

An improper phase difference may be detected with an oscilloscope. This procedure is explained in chapter 3 of

Aviation Electronics Technician 3 and 2, NavPers 10317, under the heading Frequency Comparison.

Phase adjustments are generally made with the system at rest (no input rotation). In order to provide sufficient output from the control amplifier for comparison with the input to the fixed voltage winding it is desirable to physically hold the servomotor away from its correspondence position. The phase adjustment is then made so as to provide a 90-degree phase displacement.

When a new system is installed or components replaced, it is necessary to check for proper polarity relationships. Unless the fixed and controlled voltages have the proper polarity relationships, the servo will not have stability and the motor will run away. Likewise, if a rate generator (damping signal generator, fig. 8–9) is used in the servo system, the output of the generator must have a polarity such that it will increase the systems stability rather than decrease it.

EFFECT AND ADJUSTMENT OF CONTROLLER GAIN.—The overall gain of the system has a most important effect on the servomechanism response characteristics and is one of the more easily adjustable parameters in electronic servo controllers. Increasing the system gain reduces the system velocity errors and those static errors resulting from restraining torques on the servo load or misalinement in the system. An increase in system gain also increases the system's natural frequency, and therefore its speed of response to transient inputs. However, excessive gain always decreases the rate at which oscillatory transients disappear, and continued increase in the system gain eventually produces instability. The optimum gain setting depends on the particular application for which the servo system is intended. The gain should always be as high as is commensurate with system damping.

When feedback networks are used, the relative gains of these networks and the amplifiers forward gain must be adjusted in order to obtain optimum performance. If the gain settings for these two adjustments are not given in the service instructions, they may be obtained experimentally by gradually adjusting first one and then the other.

Since the gain of the servo will be affected by the phase of the various voltages, it is desirable to check the phase adjustments before making the gain adjustments. When small servomotors are used, it is possible to measure the dynamic performance by checking the number of overshoots. This may be done by causing the rotation to cease (example—stop the antenna rotation of a radar installation), then manually rotating the servomotor a certain number of degrees from correspondence, and then allowing it to snap back to its correspondence position. The number of overshoots for a particular amount of servo displacement must be determined experimentally from equipment that is functioning properly. This may be used as a comparison guide between the same types of equipment. As a rule in aviation equipment, due to weight limitations, the servomotor has little torque in excess of that required of it. Because of this, whenever any additional load is added to the motor, the system is apt to become sluggish or intermittent. Oftentimes the servo amplifier is thought to be the cause of the above action when in reality it is due to an increase in load. This increase may be caused by excessive mechanical friction due to a worn gear train or defective bearings and may frequently be detected by manually rotating the servomotor with the power removed. The amount of effort required to rotate the system that is not operating properly is compared with that required to rotate a system that is operating properly.

BALANCE ADJUSTMENT.—Frequently servo amplifiers contain a balance adjustment to overcome any electrical unbalance that may occur in the amplifier. This unbalance is primarily caused by the unequal conduction of tubes. The adjustment is generally provided by connecting a potentiometer in the cathode circuits of the output tubes as shown by *R*21 in figure 8–7. This potentiometer is sometimes placed in the cathodes of the driver stage instead of the output stage. As the potentiometer is changed, the cathode bias will increase on one tube and decrease on the other depending upon the direction of movement of the potentiometer arm. Thus by proper positioning of the rotor, the tubes can be made to draw equal current.

When this adjustment is made it is necessary to short out the input to the servo amplifier (this is possible in some equipment by rotating the gain adjustment to the zero position). With zero input to the amplifier, the balancing adjustment is rotated to the null or minimum voltage output from the servo amplifier. This indication is taken from a meter that is connected across the output winding or from a test jack.

APPLICATIONS OF TYPICAL SERVOMECHANISMS

The remainder of this chapter will be devoted to an explanation of the theory and operation of three different servomechanisms that are typical of airborne servo systems. The first explanation will deal with a servo system used to control a loop antenna in a radio compass, the second will explain the servomechanisms used to provide azimuth stabilization, and the third a system for tilt stabilizing an airborne radar.

Radio Compass Loop Control Circuit

Essentially, the radio compass, also known as an automatic direction finder (ADF), is an ordinary radio direction finder, with the addition of automatic controls that cause the loop antenna to rotate and seek the null point, stopping there and remaining at the null point. It is able to differentiate between the correct null and the reciprocal. Also, the angular position of the loop is generally transmitted by means of synchros to an indicator for observation.

For means of explanation, the AN/ARN–6 automatic radio compass will be described. This radio compass system provides a means of automatically indicating the relative bearing of a transmitting station that has been selected by the operator. Once the operator has chosen the station on which the bearing is desired, the following procedure is followed: The operator tunes the compass receiver to the station on which a bearing is desired. The received signal from the transmitting station is picked up by the loop antenna and the nondirectional (sense) antenna. The automatic function of the control circuits causes the loop to rotate until the signal voltage induced into the loop is at a minimum point, or null. By means of synchros, the rotation of the loop antenna causes an indicator needle to turn with it, indicating the angular position of the loop at any time. When the loop has reached the null point, the indicator needle shows the relative bearing of the transmitting station with respect to the aircraft.

LOOP CHARACTERISTICS.—In order to understand how this system is able to differentiate between the correct null and the reciprocal, a brief review of loop antenna characteristics will be given.

When the plane of a loop receiving antenna is perpendicular to the direction of travel of the RF wave, the voltages induced in the two vertical sides are of equal magnitude and have the same phase. Since the voltages are directed around the loop in opposite directions, they cancel each other and result in zero loop current. On the other hand, when the plane of the loop is parallel to the direction of wave travel, the wave reaches the two sides at slightly different times, causing a phase difference that gives rise to a resultant or differential voltage acting around the loop. This is shown by the figure eight loop response curve in figure 8–11.

When a loop antenna is rotated and a measurement of the signal voltage produced by it is made, it will be found that for 360 degrees of revolution there will be two points of minimum signal voltage produced, 180 degrees apart in azimuth. Also, there will be two points of maximum

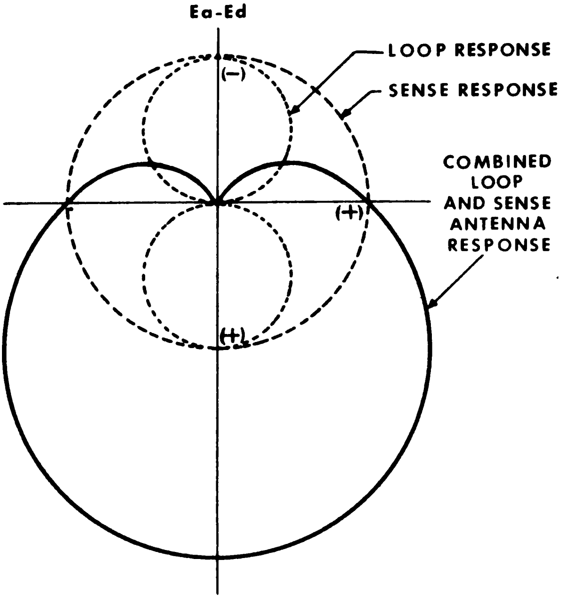


Figure 8–11.—Patterns obtained by combining sense voltage *Ea* and differential (loop) voltage *Ed*.

signal voltage produced 180 degrees apart in azimuth. The points of minimum and maximum are displaced by 90 degrees. These conditions are illustrated graphically in figure 8–11. The null points are generally used for the purpose of taking bearings for they are sharper than the points of maximum signal.

There will also be a 180-degree phase difference in the loop output voltages as measured on either side of a null point. To say in another way, this means that as the loop is turned past one of the null points, the signal output of the loop is shifted in phase by 180 degrees. This is an important point to remember in understanding the operation of the automatic compass. It is also important to point out that the phase relationship between the signal voltage produced by the loop antenna and the signal voltage produced by the nondirectional or sense antenna is such that the voltages are 90 degrees out of phase.

The magnetic field of the radio wave is assumed to be varying in space and time according to the sine curve (fig. 8–12 (A)). Assume the loop antenna to be oriented for maximum output voltage. As the magnetic flux lines of the radio wave cut the loop, a voltage of sine waveform is induced in the loop. The instantaneous values of this voltage depend on the rate of change of flux linking the loop. At the instant the flux is maximum, its rate of change is zero. Hence, the loop voltage is zero. A quarter cycle (90 degrees) later, the flux is going through zero and changing at its maximum rate. Therefore, the loop voltage is maximum.

The induced voltage in the loop, like the induced voltage in a transformer, obeys Lenz's law which states that all induced voltages oppose the force that produces them. Thus, in the first quarter cycle (0 degree to 90 degrees) the field is rising in a positive direction and the induced voltage is negative, opposing the rise. In the second

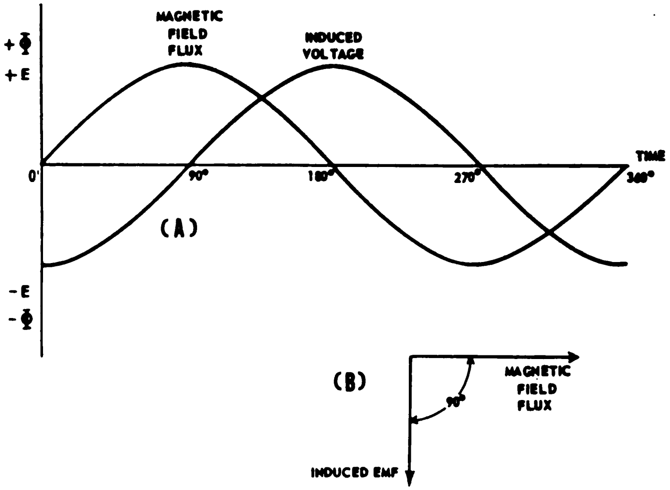


Figure 8–12.—Relation between flux and induced voltage. (A) Sine curves; (B) vectors.

quadrant (90 degrees to 180 degrees) the flux is falling and has a positive polarity, whereas the induced voltage in the loop is seen to lag the magnetic field component of the radio wave by 90 degrees (fig. 8–12 (B)).

A simple vertical (sense) antenna, as shown in figure 8–13, has no directional effect. The induced voltage in the antenna is caused by the electric component of the transmitted wave and is in phase with the electric field. When the electric field is maximum, the induced voltage is maximum. The two components of the transmitted wave, the electric and magnetic fields, are in phase. Hence, the induced voltage in the antenna is in phase with the magnetic field component. This is not true of the voltage induced in the loop by the magnetic component.

The induced voltage in the loop caused by the magnetic field is 90 degrees out of phase with the magnetic field. Comparing the voltage induced in the loop and that induced in the simple vertical antenna shows that the induced voltage in the loop is 90 degrees out of phase with the induced voltage in the simple vertical antenna. Whether the sense antenna voltage leads or lags the other is unimportant.

In the compass system provision is made to cause the loop output signal to be shifted by 90 degrees. This causes the loop output voltage to either be in phase with

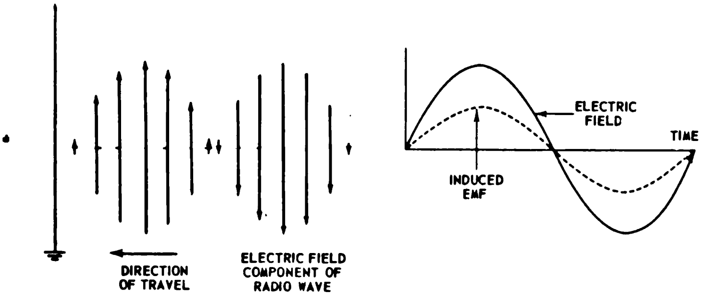


Figure 8–13.—Voltage induced in a vertical antenna by electric field component of a radio wave.

the sense antenna voltage or 180 degrees out of phase. If the loop and sense antenna voltages are combined, there will be only one null point for 360 degrees of rotation of the loop. The effect of eliminating one of the nulls produces what is known as unilateral indication (as compared to bilateral when two nulls are obtained). This is illustrated in figure 8–11.

CIRCUIT ANALYSIS.—In order to gain a fuller understanding of this circuit, make frequent reference to figure 8–14. This shows that the signal voltage induced into the sense antenna is fed into the RF amplifiers and on through the receiver's IF and audio stages. The signal voltage induced into the loop antenna is first applied to the loop amplifier to increase its signal amplitude. It is then shifted in phase by 90 degrees in the phase shifting circuit and is then applied in phase to both grids, push-push, of the balanced modulator circuit. At the same time the 100-cycle supply voltage is also fed to the balanced modulator grids, push-pull. The voltage obtained in the output of the balanced modulator is combined with the voltage from the sense antenna. These combined signals are passed down the line through the receiver.

The audio output of the last stage contains the audio modulation that was originally present on the received carrier signal and the 100-cycle component added by the balanced modulator circuit. These two components are separated and only the 100-cycle modulation component goes to the control signal amplifier. This signal is amplified and fed to the grids of the loop control tubes, a pair of thyratrons. The 100-cycle supply voltage is applied to the plates of the thyratrons. This is the same as the 100-cycle supply voltage applied to the balanced modulator circuit. Only the thyratron which has both its plate and grid driven positive at the same time will fire. The ionization and conduction of one of the thyratrons cause current to flow through one of the field windings of the loop drive motor, causing it to rotate. The direction of

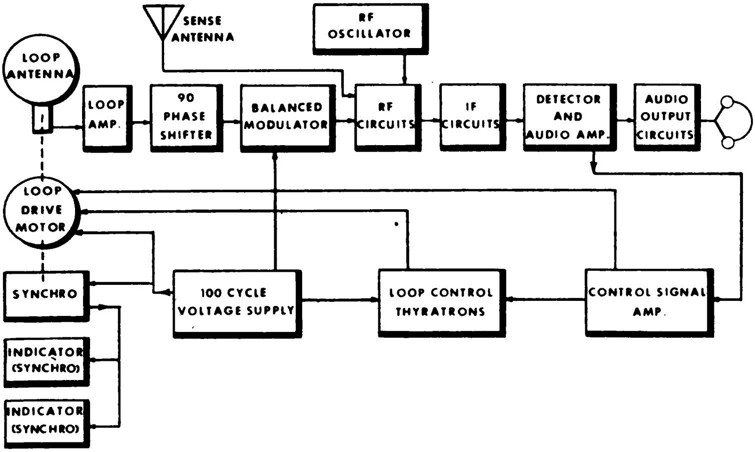


Figure 8–14.—Loop control circuits, AN/ARN–6 radio compass.

current flow, and thus the direction of rotation, is determined by which thyratron is fired. The phase relationship of the grid and plate signals determine which tube shall fire. This phase relationship is determined by the direction from which the signal is arriving. The net result is that the loop is rotated until it reaches a null point, at which time rotation will stop for lack of a signal.

The turning of the loop antenna causes the rotor of a synchro generator to turn since they are mechanically coupled. The rotor of the synchro motor, which serves in the indicator, will also turn due to the changes in its stator voltages, since the stators of both the synchro generator and motor are electrically coupled. The synchro motor drives the indicator pointer, thus indicating the bearing of the station.

In order to better understand the following circuit action you should make frequent reference to the schematic shown in figure 8–15: A SIGNAL VOLTAGE is induced into the loop antenna when the receiver has been tuned to the desired station. This loop signal voltage is transformer coupled into an RF amplifier circuit which amplifies the signal and feeds it on to the phase shifting circuit.

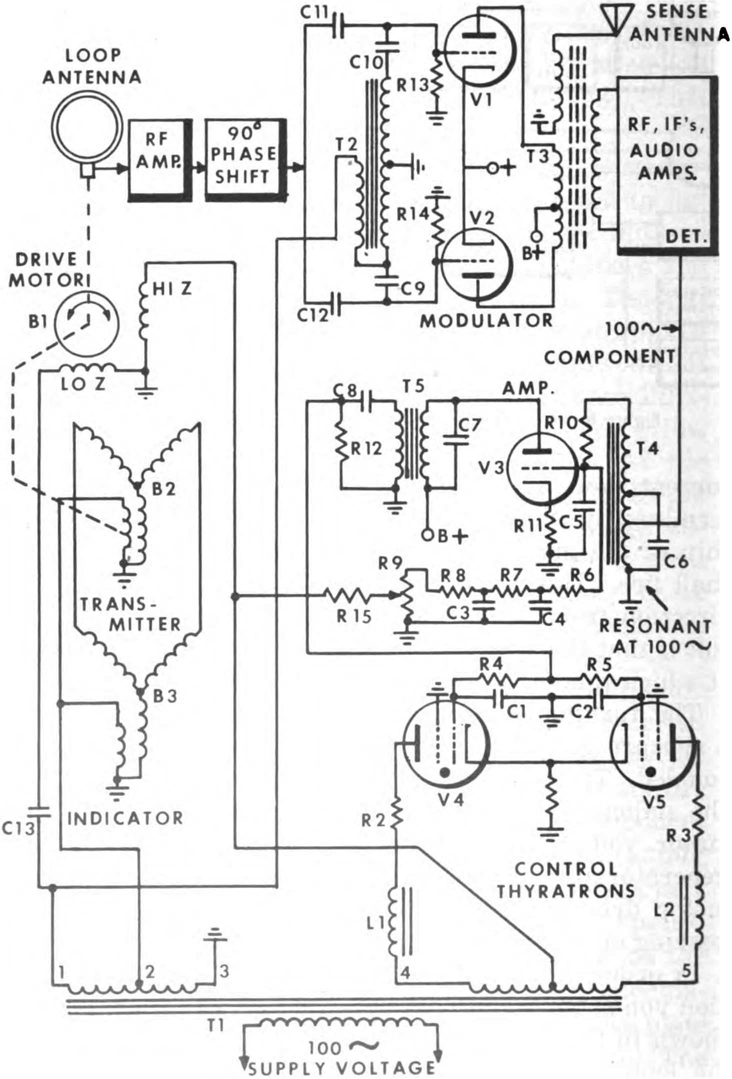


Figure 8–15.—Simplified schematic—loop control circuits.

(Phase shifting circuits are discussed in chapter 2.) This part of the circuit is shown in block form on the diagram since its operation is not essential to understanding the rest of the system. The phase shifting circuit shifts the phase of the signal voltage by 90 degrees before feeding it into the modulator circuit. Whether or not the shift is ahead 90 degrees or back 90 degrees is unimportant. In either case, the shift causes the voltage to be either in phase with or 180 degrees out of phase with the sense antenna voltage.

The loop voltage, after being shifted in phase by 90 degrees, is applied to the MODULATOR CIRCUIT through coupling capacitors *C*11 and *C*12 to the grids of the modulator tubes. Since the signal is coupled from the same point to both grids, they are fed in phase, or push-push. Tubes *V*1 and *V*2 serve as amplifiers, the plates of both tying into the primary winding of *T*3, which serves as the RF input transformer for the RF circuits of the receiver. Since the plates of *V*1 and *V*2 feed opposite ends of the primary of *T*3, and since the grids of the two tubes are fed in-phase signals, it might appear that no voltage will be induced into the secondary. But in addition to the loop signal voltage being fed in phase to the two grids, there is also another voltage applied to them. A 100-cycle voltage is taken from pin 1 of the winding of *T*1 and applied to one end of the autotransformer *T*2. The 100-cycle voltage induced into the secondary end of *T*2 is applied out of phase to the two grids of the modulator circuit, or push-pull. The effect of the 100-cycle voltage is to cause *V*1 and *V*2 to conduct alternately, so that the loop signal on the grids is amplified first by one tube and then by the other. In other words, only one tube will be allowed to amplify the loop voltage during each half cycle of the 100-cycle modulating voltage. This is equivalent to shifting the signal input from one end of the primary of *T*3 to the other, or, in other words, reversing the phase of the signal.

By referring to figure 8–16, the following may be

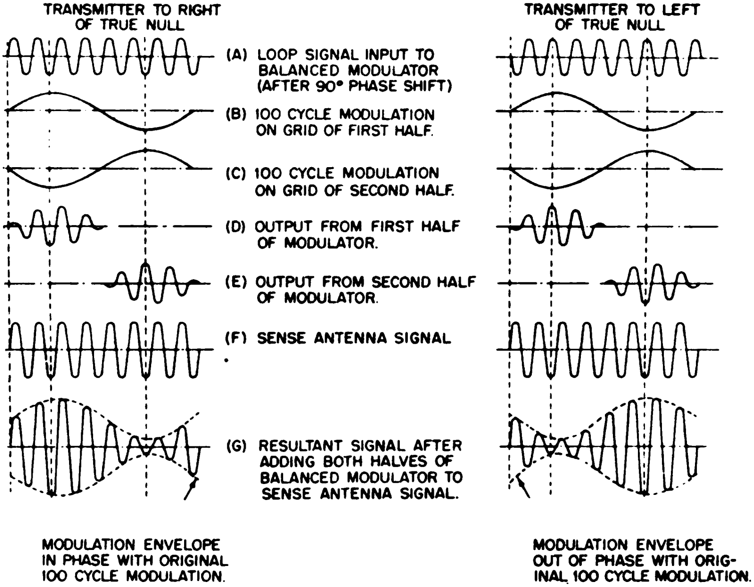

TRANSMITTER TO RIGHT OF TRUE NULL
TRANSMITTER TO LEFT OF TRUE NULL
(A) LOOP SIGNAL INPUT TO BALANCED MODULATOR (AFTER 90° PHASE SHIFT)
(B) 100 CYCLE MODULATION ON GRID OF FIRST HALF.
(C) 100 CYCLE MODULATION ON GRID OF SECOND HALF.
(D) OUTPUT FROM FIRST HALF OF MODULATOR.
(E) OUTPUT FROM SECOND HALF OF MODULATOR.
(F) SENSE ANTENNA SIGNAL
(G) RESULTANT SIGNAL AFTER ADDING BOTH HALVES OF BALANCED MODULATOR TO SENSE ANTENNA SIGNAL.
MODULATION ENVELOPE IN PHASE WITH ORIGINAL 100 CYCLE MODULATION.
MODULATION ENVELOPE OUT OF PHASE WITH ORIGINAL 100 CYCLE MODULATION.

Figure 8–16.—Balanced modulator input and output.

noted: The loop signal applied to the grids of the modulator after being shifted 90 degrees by the phaser circuit is shown at (A). When the transmitter is to the right of the true null, the loop signal to the balanced modulator is shown at the left of the figure. When the transmitter is to the left of the true null, the signal is as shown at the right of the figure. Note that there is a phase difference of 180 degrees. At the same time that one of these signals is being applied in phase (push-push) to the grids of *V*1 and *V*2, the 100-cycle modulating voltage is being applied out of phase (push-pull) to the grids.

The voltage on one grid is shown at (B) and on the other grid at (C). The result in the plate circuits will be as shown at (D) and (E). These outputs are present in the primary of *T*3. The signal voltage induced into the nondirectional (sense) antenna is fed to another primary winding of *T*3; this sense antenna signal voltage is shown at (F). These two signals are combined in the secondary of *T*3, resulting in the waveforms shown at (G). Notice that the envelope at the left side of the figure is in phase with the original 100-cycle modulating voltage as shown at (B) and (C), while, when the transmitter is to the opposite side of the null the modulation envelope is out of phase with the 100-cycle modulating signal as shown at the right side of the figure. In other words, after this resultant signal has passed through the receiver and been demodulated, the resulting 100-cycle audio component of the output has a phase relationship with respect to the original 100-cycle modulation voltage which is dependent upon the direction of signal reception with respect to the position of the loop. The combined sense antenna and modulated loop antenna voltage pass on through the compass receiver.

The remainder of the compass receiver is shown in block form since its operation is conventional and has no bearing upon the topic being presented. The combined voltages mentioned are amplified and detected. The audio component that is detected from the received signal is passed to the audio amplifier circuits. The 100-cycle component produced by the modulator is removed from the first audio stage and passed to the control signal amplifier *V*3. This 100-cycle component is the only portion with which we are concerned.

A portion of *T*4, in the grid circuit of the CONTROL SIGNAL AMPLIFIER, *V*3, is shunted by *C*6 to form a parallel resonant circuit that is resonant at 100 cycles. The 100-cycle signal input is the control signal for the loop control circuits. It may be noted at this point that the receiver circuits are so designed that for automatic operation of the loop the receiver is not able to produce a 100-cycle beat note for CW reception, as an additional 100-cycle signal would affect operation of the loop control circuits. The control circuit amplifier is a fairly conventional class A amplifier, except for the R–C network connected to the grid which is used to prevent HUNTING and is explained later. The plate circuit tank is tuned to 100 cycles. The 100-cycle voltage produced into the secondary of *T*5 is fed to the grids of the control thyratrons.

The CONTROL THYRATRONS are the loop control tubes in that they directly cause the loop drive motor to operate. The 100-cycle control voltage from *T*5 is applied to the grids of *V*4 and *V*5, the thyratrons, in phase (push-push). The plates of *V*4 and *V*5 are supplied with a 100-cycle voltage from pins 4 and 5 of *T*1, which also supply the 100-cycle voltage for the modulator. Since the plates of *V*4 and *V*5 are tied to opposite ends of the secondary winding, they are fed out of phase with each other, or push-pull. IMPORTANT: Recall that the 100-cycle control voltage coming from *T*5 will either be in phase with or 180 degrees out of phase with the original 100-cycle modulation voltage depending upon the loop position. Since one plate is positive and the other negative at any given moment, only the tube which has both its plate and grid positive at that moment will fire. This is determined by the phase relationship of grid and plate voltage as discussed previously. This means that with the loop in one position, *V*4 may have the proper phase relationship of voltages on its grid and plate to permit it to conduct. If the loop is rotated past a null point, *V*5 will then have the in-phase voltages on its grid and plate and it will conduct.

The conduction of either *V*4 or *V*5 will cause current to flow in the high impedance (HI Z) winding of the loop drive motor, *B*1. The low impedance (LO Z) winding of the drive motor is supplied from a winding on *T*1, and each half cycle the current flow through the winding reverses. The phase relationship of the current in the high and low impedance windings depends upon which of the thyratrons is conducting, and this in turn determines the direction of motor rotation. It should be noted that current flows in the high impedance winding for only one-half of each cycle of the 100-cycle supply voltage. This is true since the controlling thyratron conducts only when its plate is positive and neither thyratron conducts during negative half cycle of the grid signals. The motor will rotate and the loop will turn until the null point is reached, at which time the motor will no longer be driven because of the lack of a signal being induced into the loop.

An ANTIHUNT FEEDBACK ARRANGEMENT is provided in this equipment to prevent hunting. This is necessary because, due to inertia, the loop drive assembly will overshoot its null position. After passing the null, the assembly will automatically return toward the null and overshoot in the opposite direction. This action of the loop moving back and forth across the null point is called HUNTING. A negative feedback arrangement is used to minimize this action.

The high impedance winding of the loop drive motor is controlled by the thyrathrons. The voltage fed to this winding is also applied to the voltage dividing system of *R*8 and *R*9. A portion of this voltage, which is determined by the setting of *R*9, is attenuated and filtered by the R–C network of *R*8, *C*3, *R*7, *C*4, *R*6, and *C*5. The signal appearing at the output of this network is applied to the grid of *V*3, along with the original 100-cycle control signal from the audio output circuit of the receiver. The phase of the feedback signal will always be opposite to the original 100-cycle signal from the audio circuit and will tend to counteract it.

Until the loop approaches the null point, the feedback voltage will be too small in comparison to the original 100-cycle voltage to have any effect on loop rotation. When the loop approaches the null point, the 100-cycle signal from the audio circuit decreases in amplitude until the feedback is, by comparison, now large enough to exercise some control on *V*3. Thus, as the loop reaches the null point, its speed will have been reduced by the feedback signal so that it will not swing past the null position. The potentiometer *R*9, sometimes called automatic sensitivity adjustment, may be adjusted so that the correct amount of feedback signal is present on the grid of *V*3 and hunting will be at a minimum. In general, if enough voltage is not fed back, the loop oscillates slightly about the null. If too much is fed back, the loop rotates sluggishly in the vicinity of the null.

The angular position of the loop is transmitted by means of a SYNCHRO SYSTEM to one or more indicators for REMOTE INDICATION. The rotor of a synchro generator is mechanically connected to the loop assembly. The stators of this synchro are connected to the stators of a synchro motor in the indicator units. The 100-cycle supply furnishes the excitation voltage for the operation of the synchro units and is applied to all rotors. When the loop is turned, the rotor of the synchro generator turns, changing the voltages induced into its stator. This changes the values of voltage appearing on the stators of the synchro motor in the indicators. The rotors of the synchro motors turn to the new position. The indicator pointer is tied to the rotor of the synchro motor and indicates the new heading of the station being received.

Azimuth Stabilization

Most azimuth stabilization systems are similar, and an understanding of one system will provide information that will be helpful in the comprehension of other systems. For purposes of explanation, the azimuth stabilization system of the AN/APS–33 radar will be discussed in this chapter.

An azimuth stabilization system stabilizes the indicator screen presentation causing the display to remain rotationally fixed, despite changes in the heading of the aircraft. By this means, a definite compass relationship is established for the indicator screen, and radar target echoes appear in proper position with respect to the top center of the screen, which represents magnetic north. The stability of the indicator presention also facilitates search during rough weather or moderate maneuvering of the aircraft.

Stabilization of the indicator screen pattern is accomplished by signal voltages controlled by the master indicator of the fluxgate compass in the aircraft. These signals are used to control the angle of the sweep trace on the screen in accordance with the instantaneous azimuth directivity of the scanner.

In order to explain the theory of operation of azimuth stabilization it is necessary to analyze briefly one of the compass systems used. This is necessary since in an equipment that contains azimuth stabilization the compass that controls the stabilization is an intricate part of the system. The GYRO FLUXGATE COMPASS SYSTEM of the aircraft is a remote-indicating earth-inductor compass, which consists of a gyro-stabilized fluxgate transmitter, an amplifier, a master indicator, and one or more synchro remote-indicator repeaters. (NOTE: The term "gyro fluxgate" is a registered trade name of the Bendix Corporation.) The fluxgate element is stabilized in the horizontal plane by an electrically-driven horizon gyro. The gyro head contains a rolling-ball type of erecting system, which maintains the gyro axis in the vertical plane by setting up a restoring force when the axis becomes displaced.

The transmitter portion contains the sensitive element of the system, the fluxgate unit, which consist of three equal legs forming a triangle; this unit is established and maintained in the horizontal plane by the action of the gyro. Each leg consists of two strips of a magnetic alloy used as a core upon which are wound a primary and a secondary coil. The primary coils are connected in series and are excited by an oscillator circuit in the amplifier. The secondary coils are delta-connected, with leads tapped off at the points where the three coils join. These leads, and the single lead which supplies the excitation current, are brought out through hairsprings on the gimbal rings. The primary coil of each leg is wound in such a manner that when the leg is at right angles to the magnetic field of the earth the exciting alternating current produces opposing fluxes in each strip which neutralize each other and induce no voltage in the secondary.

This exciting current saturates the core on the peak of each half cycle.

When the leg is parallel to the earth's magnetic field, the earth's flux tends to oppose the flux in one alloy strip and to aid the flux in the second strip. The partial cancellation of the flux in the first strip prevents its reaching saturation at the same time as the second strip, and its flux continues to change after the second strip has reached saturation. The flux difference thus set up induces a voltage in the secondary. When the current through the primary decreases to a point where the flux starts to drop below the point of saturation, a voltage in the opposite direction is induced in the secondary. This occurs because the flux in the first strip (in which partial cancellation exists) drops below the saturation point before the flux in the second strip, thus producing a flux change in the opposite direction. Since saturation occurs twice each cycle, and since secondary voltages are induced at the start and again at the finish of each period of saturation, the frequency of the induced voltage is double that of the primary voltage.

Three signal voltages, which generally are of three different magnitudes, are obtained from the fluxgate secondary coils, and are applied to the stator coils of a coupling synchro contained in the master indicator. The voltage induced in the rotor winding of this synchro is fed to the fluxgate amplifier, which develops sufficient power to operate a 2-phase induction motor which is also in the master indicator. The fluxgate amplifier includes an extremely stable oscillator that is used to furnish excitation power for the fluxgate, and another oscillator which provides voltage for the continuously excited winding of the 2-phase induction motor in the master indicator. The excitation voltage applied to the induction motor is necessarily at a frequency of twice that of the fluxgate excitation voltage in order to equal the frequency of the fluxgate output.

The 2-phase induction motor in the master-indicator phase unit has its stator wound in quadrature (two windings at right angles to each other). The control winding is energized by the amplified signal from the synchro rotor, and the exciting winding receives its power from the oscillator in the amplifier. When both windings of the induction motor are energized by voltages having a phase difference of 90 degrees, the motor rotates and drives the synchro rotor through a gear train until the null position is reached. When this occurs the voltage induced in the rotor drops to zero. The output of the amplifier also becomes zero, and since only the excitation winding of the induction motor is energized, the motor stops. The position of the synchro rotor is therefore determined by the position of the fluxgate with respect to the earth's magnetic field; consequently, when the aircraft turns, changing the azimuth relation of the fluxgate to the earth's magnetic field, the synchro rotor and attached indicator pointer are turned by the induction motor through the same angle and in the same direction. When the synchro rotor reaches the null position, the master indicator shows the magnetic heading of the aircraft. A transmitting magnesyn, which is also turned by the induction motor, provides controlling voltages for remote-indicating repeaters.

ANALYSIS OF AZIMUTH STABILIZATION.—A simplified schematic of the azimuth stabilization section of the AN/APS–33 radar that will be used in this analysis is shown in figure 8–17. The MASTER INDICATOR of the fluxgate compass system contains the transmitting synchro used with this azimuth stabilization system. The rotor of the transmitting synchro is mechanically coupled to the shaft that drives the pointer of the master indicator. As the aircraft changes its heading, the rotor correspondingly changes position at a one to one ratio. The stator of the transmitting synchro is electrically coupled to the contacts of relay *K*2101. This relay is energized by the operation of the bearing switch on the control unit of the radar system. By referring to figure

8–17, it can be seen that *K*2101 must be energized before the output of the transmitting synchro can be utilized in stabilization.

The ZERO SET SYNCHRO is utilized to zero set the PPI sweep position when the output of the transmitting synchro is not used, which is during relative bearing operation. The stabilization system will first be described as it relates to relative bearing operation; later it will be shown how azimuth stabilization is provided when the transmitting synchro of the aircraft compass system is connected to the azimuth followup synchro. When on relative bearing operation, the indicators are nonstabilized and the top of the indicator represents the heading of the aircraft; the targets are displayed with respect to the heading of the aircraft. The rotors of the zero set synchro and the transmitting synchro of the fluxgate compass are supplied excitation voltage from the same source. The zero set synchro is adjusted manually. Its positioning will be discussed later.

The stator of the AZIMUTH FOLLOWUP SYNCHRO is electrically coupled to either the stator of the transmitting synchro in the fluxgate compass, or the stator of the zero set synchro. The stator that is used for coupling depends upon the position of *K*2101. The rotor of the followup synchro is mechanically connected to the azimuth torque motor and the secondaries of the azimuth sweep synchro. The phase relationship between the rotor of the transmitting synchro or the zero set synchro determines the amplitude and polarity of the signal developed at the rotor of the followup synchro. When, for example, the rotors of the followup and the zero set synchros are 90 degrees displaced, as shown in figure 8–17, there will be no signal at the rotor of the followup synchro. If, however, the zero set rotor is turned 10 degrees, an error signal will be induced into the rotor of the azimuth followup synchro. The amplitude of the error signal produced will be dependent upon the amount of displacement between the rotors. The polarity or phase of the signal will depend upon the direction of displacement. The error signal is fed to the servo amplifier input of the true bearing amplifier which drives the azimuth torque motor.

The **AZIMUTH TORQUE MOTOR** is a 2-phase induction-

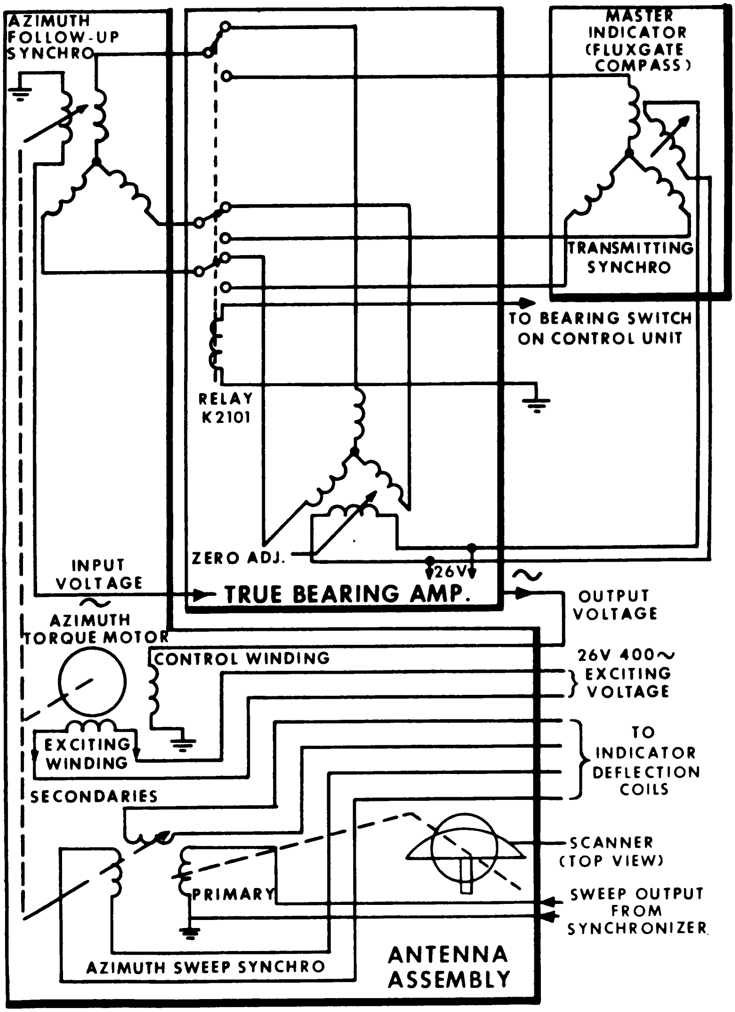


Figure 8–17.—Analysis of azimuth stabilization.

type motor. The fixed voltage winding of this motor is connected to the same power source that supplies the rotors of the transmitting and the zero set synchros. The control winding is connected to the output of the true bearing amplifier. The amplitude and phase of the error signal applied to the control winding depend upon the amount and direction of displacement of either of the two exciting rotors (zero set or transmitter). The azimuth torque motor turns in a direction that is determined by the phase relationship of the voltages applied to these two windings. Since the torque motor is mechanically coupled to the azimuth followup synchro rotor, as it turns, it drives the followup rotor to a new null relationship with the rotor of the zero set synchro. As the rotors once again approach their 90-degree relationship, the amplitude of the error signal decreases, thus decreasing the amplitude of the voltage applied to the control winding of the torque motor. When the rotors of the zero set and the followup synchros are once again displaced by 90 degrees, there is no longer a signal applied to the control winding and the torque motor stops. As previously stated, the azimuth torque motor also turns the secondaries of the azimuth sweep synchros; thus, shifting the position of the azimuth sweep secondaries changes the position of the sweep on the indicators. For example, assume that the antenna is in a dead ahead position, with the sweep pointing to the top of the indicator. Then, by adjusting the zero set synchro to a setting of 10 degrees, which shifts the sweep secondaries by 10 degrees, the sweep position would move to 10 degrees.

The primary (rotor) of the azimuth sweep synchro is mechanically coupled to the antenna, thus when the antenna rotates the primary winding is also rotated. The primary winding is supplied with sweep voltages from the radar synchronizer. These voltages are induced into the secondary coils with a variation of phase and amplitude for each coil, depending upon its instantaneous angular relationship with the primary. The secondary voltages are applied to the vertical and horizontal coils of the PPI indicator. By means of these voltages, the angular deflection of the sweep trace on the PPI indicator is controlled in accordance with the instantaneous azimuth directivity of the radar antenna. The example cited previously utilized the movement of the zero set synchro (manual adjustment) to move the secondaries of the azimuth sweep synchro and hence the PPI sweep. The same comparison may be made with the transmitting synchro. This synchro changes as the heading of the aircraft changes and this action coupled with the operation of the system just discussed is the analysis of azimuth stabilization.

COMPARISON OF THE TRUE AND RELATIVE BEARINGS.—As already explained, changing the aircraft heading changes the positions of the sweep secondaries of the azimuth sweep synchro. This occurs on azimuth stabilization only and the change is automatic. On manual operation the only change that can be made is by varying the position of the zero set synchro. This is not stabilization, for it has no relationship to the movement of the aircraft. The zero set synchro is adjusted on manual or relative bearing operation. The antenna is set on zero degree with respect to the aircraft and the zero set synchro is adjusted until the sweep appears at the dead ahead position.

During relative bearing operation, with the azimuth indicating device at the top of the scope, and targets appearing thereon, a change in the bearing of the aircraft will result in a change of position of targets. This is an undesirable feature due to the fact that during the time the targets are shifting position they are hard to observe on the indicators. See figure 8–18 (A) for a visual explanation of this condition. The top of the scope indicates the heading of the aircraft and all targets are displayed relative to this heading, thus on relative bearing operation the top of the indicator indicates the heading of the aircraft. In other words, when the

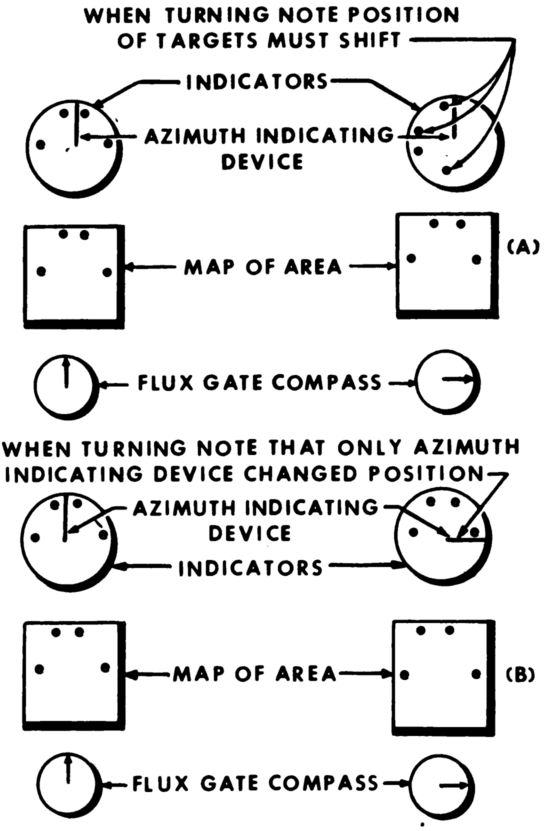


Figure 8–18.—Comparison of relative and true bearing operation.

aircraft is heading north, the top of the scope is north. When the aircraft is heading in any particular direction, the top of the scope is also that direction.

The fluxgate compass controls the position of the sweep on the indicators during true bearing operation. As the fluxgate compass follows the aircraft's changes in heading, the error signals coupled to the azimuth stabilization system change the position of the sweep. These changes in position occur at a one to one ratio. Thus, if the aircraft heading changes a certain number of degrees, the sweep secondaries will also change a like amount. Refer to figure 8–18 (B) for a comparative analysis of true bearing operation.

In this operation the azimuth indicating device, as on relative operation, represents the heading of the aircraft (providing the marker control on the AN/APS–33 control box is set to zero degrees). The action of moving the sweep secondaries to correspond with changes in the heading of the aircraft causes the azimuth indicating device to shift. Hence, the azimuth indicating device will always indicate the heading of the aircraft and, when on true bearing operation, the top of the scope always indicates north (providing the marker control on the AN/APS–33 control box is set to zero degrees). As the aircraft changes its heading, the targets remain in their position on the scope and only the azimuth indicator moves. Thus, the targets remain fixed, are always visible, and their relationship to the aircraft can be checked with the azimuth indicating device.

The top of the indicator represents magnetic north when the VARIATION is not set in, and true north when the variation has been set in. Variation is the angle between magnetic north and true north and differs at different points on the earth. The number of degrees of variation may be obtained from navigational charts. Some compass systems provide a means whereby variation may be added into the system, thus the compass will then indicate true north and a radar slaved to the compass will also indicate true north. In some aircraft installations that are equipped with a compass that does not provide this correction feature, a variation differential generator is included. This generator is connected between the transmitting synchro of the compass and the azimuth followup synchro in the radar. By adjusting this generator to the local variation it is possible to correct the radar to true north.

TRUE BEARING AMPLIFIER.—(Refer to fig. 8–19.) The input to the true bearing (servo) amplifier is obtained from the rotor of the azimuth followup synchro. This error signal is applied across resistor *R*1 (sensitivity adjustment) which is paralleled by the primary of input transformer *T*1. The amplitude of this signal is determined by the amount of change in the aircraft heading, and the phase is determined by the direction of the change. The signal is transformer coupled to the grids of *V*1 which is the true bearing amplifier discriminator. The signals on the grids of *V*1, due to the transformer action, will always be 180 degrees out of phase. The plate voltages, due to the transformer connections, are always in phase. Thus, the grid that is in phase with the plate will determine which half of the duodiode conducts the hardest.

When the grid and plate of *V*1(*A*) are in phase, its conduction charges the capacitors in the plate circuit. The plate current circuit is from the plate of *V*1(*A*) through the source potential winding, plus to minus, then up through resistor *R*5. The current flowing through *R*5 develops a negative voltage at the bottom and charges capacitor *C*6. *C*6 in conjunction with *C*5 provides an a-c filter path to ground. The current flows from the junction point of resistors *R*4 and *R*5 back to the center tap of transformer *T*1. The path is then completed back to the cathode through the resistor *R*2 and the balance pot *R*3. During this same time, although *V*1(*B*) has its grid negative and plate positive, it may still conduct a little. Its conduction path from the plate through the source potential causes current to flow through resistor *R*4 and develop a negative voltage at the top of *R*4. The negative voltage developed at the bottom of *R*5, however, far exceeds the negative voltage developed at the top of *R*4. Thus the voltage measurement from *R*5 is negative to positive. The voltage thus developed is coupled to the grids of the output control tubes.

The grid of *V*2(*B*) is more negative than the grid of *V*2(*A*). Any portion of the a-c signal not filtered out through capacitors *C*4, *C*5, and *C*6 is coupled through *C*7 and *C*8 to the grids of *V*2. In addition, the d-c signal

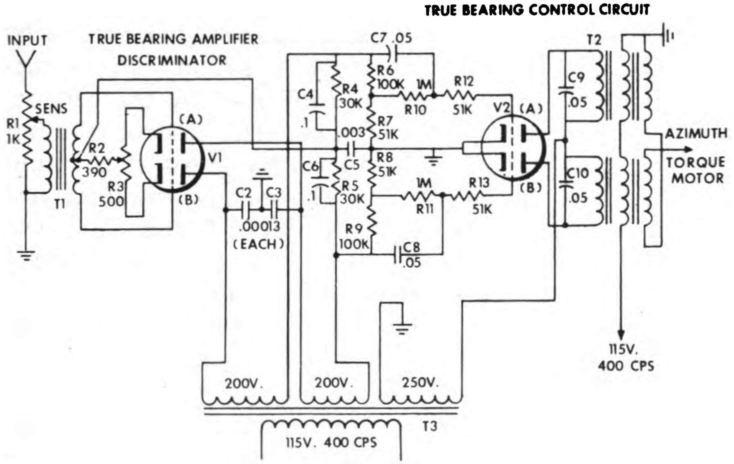


Figure 8–19.—AN/APS–33 true bearing amplifier.

developed across *R*6, *R*7, *R*8, and *R*9 is taken from the junction of the resistors and also applied to the grids. Hence, the grids contain both a-c and d-c components.

When the amplitude of the signal on either grid is too great, grid current is drawn; this limits the conduction of the tube. For example, when the amplitude of the signal on the grid of *V*2(*A*) is excessively positive, grid current will flow through *R*10 and *R*12, thus adding to the bias on *V*2(*A*).

The cathode of *V*2 is tied directly to ground. The voltages applied to the plates of *V*2 are in phase and conduction depends upon the grid signals that are developed in the discriminator circuit. In the case cited previously where *V*1(*A*) conducted the hardest, the voltages developed in the output resistors made the grid of *V*2(*B*) most negative. Under these conditions, the current flowing through the plate windings of the bottom control winding of the SATURABLE REACTOR assembly, *T*2, will be less than the current flowing through the upper winding. The current flowing through the control windings of the saturable reactor assembly determines the amount of output signal developed. (Saturable reactors are described in detail in chapter 9.) The center winding or primary of the saturable reactor assembly is excited by 115 volts a. c. The coupling from the primary to the output or secondary windings is controlled by the conduction through the control windings. The control winding with the least current flow allows the greatest coupling from primary to secondary. The output windings are wound in series opposing. Capacitors *C*9 and *C*10 serve to filter the a-c component and give a fairly constant current flow through the control windings.

In the case cited previously, since *V*2(*B*) is conducting the least, the saturation of the bottom section of *T*2 is less than that of the top section. Under these conditions the voltages developed in the bottom output winding far exceed those developed in the top winding. The polarity of the voltage taken from the output of the saturable assembly will be the same as that from the bottom winding. This voltage is applied to the control winding of the azimuth torque motor.

The rotor of the AZIMUTH TORQUE MOTOR is mechanically connected to the secondaries of the azimuth sweep synchro and the rotor of the azimuth followup synchro. The motor then drives the followup synchro rotor to a new null relationship with the rotor of the transmitting synchro, or zero set synchro, depending upon the position of *K*2101. One secondary winding of the AZIMUTH SWEEP SYNCHRO (resolver) is connected to the vertical winding of the deflection system, and the other is connected to the horizontal deflection winding.

See figure 8–20 for a descriptive analysis of changes in the azimuth stabilization system when the aircraft heading is changed. With the antenna stopped in the dead ahead position, as shown in (A), an error signal is applied to the true bearing amplifier which causes the azimuth torque motor to turn. The sweep secondaries are then driven in a direction corresponding to the change in the transmitting synchro rotor, as shown in (B). When the aircraft changes its heading, the sweep position on the PPI scope will change accordingly.

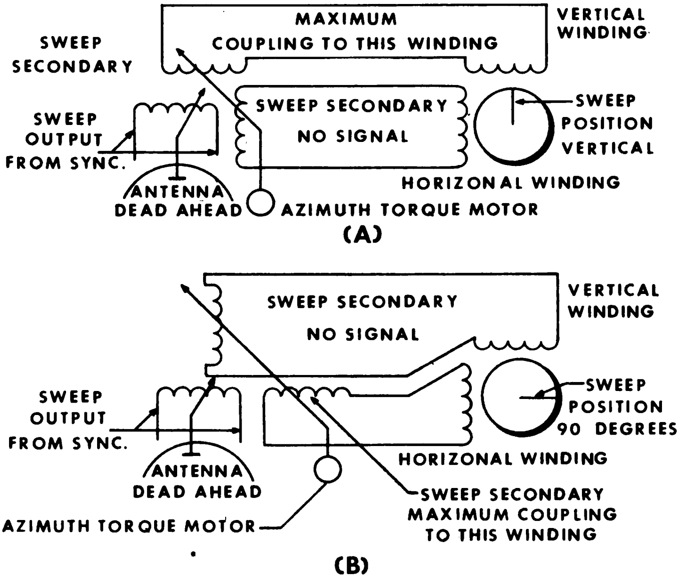


Figure 8–20.—Analysis of changes in an azimuth stabilization system.

Tilt Stabilization

The PURPOSE OF TILT STABILIZATION is to stabilize the tilt angle of the antenna of a radar system with respect to the earth. If the antenna is left unstabilized, when the aircraft changes its flight angle the antenna follows this change. If the antenna is dead ahead with respect to the aircraft and the plane begins a climb, the antenna will point up, thus losing targets. If, on the other hand, the antenna is pointing to port or starboard and the aircraft rolls to either side, the target returns will again be lost. If the antenna is stabilized with respect to the earth then the target loss, due to changes in the attitude of the aircraft, will be minimized. For example, assume that the antenna is pointing dead ahead in relation to the aircraft and the aircraft goes into a climb; if the antenna is stabilized, it will tilt down and maintain a constant angle with respect to the earth. The same holds true if the antenna is pointing to port or starboard and the aircraft is rolling; the antenna will tilt up or down, maintaining the constant reference with respect to the plane of the earth.

The stabilization range of Radar Set AN/APS–33B, which will be used for purposes of explanation, is such that if the TILT control is set for 0, the beam is maintained at this angle for aircraft pitch or roll up to plus 30 degrees or minus 30 degrees. If the TILT control is set for other beam angles, these limits vary accordingly.

The tilt-stabilization system consists of a gyroscope, servo amplifier, resolver synchro, manual-tilt synchro, tilt-torque motor, damping-signal generator, and followup synchro. (A diagram of this system is shown in figure 8–9.) The gyroscope unit is separately mounted on the aircraft structure, and the resolver synchro and the servo amplifier are mounted on the antenna-assembly frame. The tilt-torque motor, damping-signal generator, and followup synchro are contained in the tilt-servo unit mounted on the rotary section of the antenna assembly. The manual-tilt synchro is mounted in the radar set control box.

Stabilization is effected by the use of two signals, the resolver-synchro signal and the manual-tilt synchro signal. Separate pitch and roll signals originating in the gyroscope unit are combined in the resolver synchro to produce the single resolver-synchro signal. The manual-tilt synchro signal originates in the followup synchro and provides the tilt-stabilization system with continuous information as to the tilt angle of the reflector. This signal passes through the manual-tilt synchro in the radar set control box where it is modified according to the setting of the TILT control. The resolver-synchro signal and the manual-tilt synchro signal, which differ in phase by 180 degrees, are combined at the input of the servo amplifier, and the amplifier resultant, or error signal, is used to control the tilt-torque motor in correcting the tilt angle of the reflector.

GYROSCOPE UNIT.—The gyroscope unit contains the gyro, with its drive motor, erecting mechanism, motor-driven caging and uncaging mechanism, and two synchros. The gyro, powered by a 26-volt, 400-cycle, 3-phase induction motor, establishes true vertical. The caging and uncaging mechanism is electrically operated, and is manually controlled by the GYRO switch on the radar set control. When caged, the gyro is clamped in an approximately vertical position with respect to its housing. CAGING OF THE GYRO prevents it from being damaged when the aircraft is engaged in maneuvers involving pitch or roll in excess of 60 degrees. The gyro develops a normal operating speed of 18,500 r.p.m.; when it is UNCAGED, the rolling-ball type ERECTING mechanism brings the gyro to true vertical and maintains it in this position. The gyro should be uncaged when in use and caged when the equipment is turned off. Although the GYRO switch may be in the UNCAGE position, the gyro is automatically caged when the POWER switch is turned off. After the equipment is turned on, the gyro remains caged during the warmup period of approximately three minutes, after which time the operation of the caging mechanism is dependent upon the position of the GYRO switch. The schematic diagram of the gyroscope unit is shown in figure 8–21.

The PITCH and ROLL synchros in the gyroscope unit deliver signals to the tilt-stabilization system in the antenna assembly in accordance with the departure of the aircraft from level flight; these signals are used to stabilize the tilt angle of the beam with respect to the horizon. The rotor winding of each synchro is excited by 26-volt, single-phase, 400-cycle a.c. When either of the rotors is displaced from the null point, single-phase voltages are induced in the stators; the phase relationships and amplitudes of these voltages are determined by the direction and degree of displacement.

The stators of the two synchros are rigidly attached to the gyro, at right angles to each other. The rotor of the pitch synchro is mounted with its shaft horizontal,

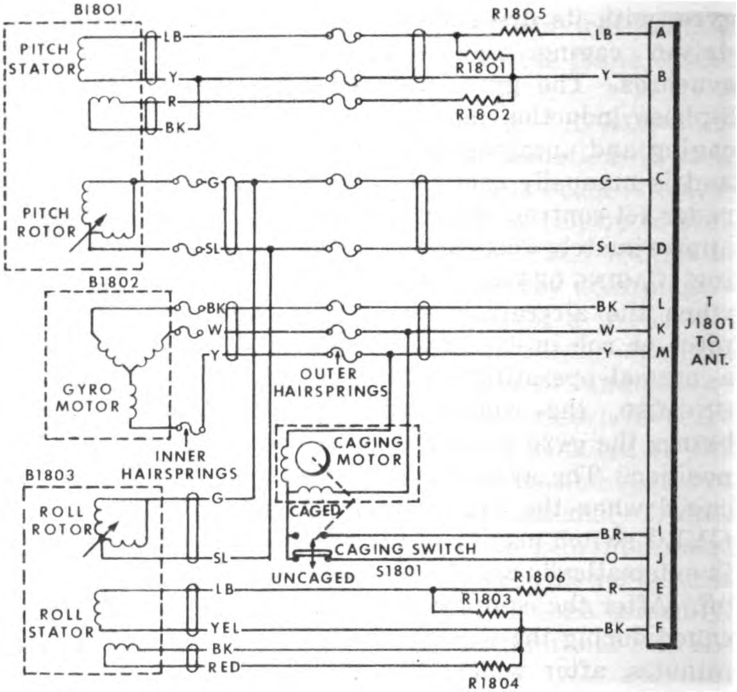


Figure 8–21.—Gyroscope unit.

and at right angles to the line of flight of the aircraft. Since the rotor is fixed with respect to the aircraft, and the stator is fixed with respect to the gyro housing, an a-c pitch signal is developed when the rotor is displaced by the pitching of the aircraft.

The rotor of the roll synchro is also mounted with its shaft horizontal, but the shaft is parallel to the line of flight of the aircraft. Since the rotor is fixed with respect to the aircraft, and the stator is fixed with respect to the gyro, an a-c roll signal is developed when the rotor is displaced by the rolling of the aircraft. When the attitude of the aircraft combines both pitch and roll, pitch and roll signals are simultaneously produced by the gyroscope unit. These signals are fed to the resolver synchro in the antenna assembly. (See fig. 8–9.)

When displacement of the gyroscope synchros occurs, a torque is developed at the stator coils, which, if not compensated for, will cause the gyro to deviate from true vertical, thereby introducing errors into the reflector angle of stabilization. To counteract this effect, a balancing coil connected to a load resistor is mounted along with the stator of each synchro. When displacement of the rotor occurs, the magnetic field developed about the balancing coil causes a torque in the direction opposite to that already exerted on the stator, thus insuring stability of the gyro. The gyroscope unit is mounted on the aircraft structure and is provided with a reference base plate which is adjusted to a plane parallel to that of the antenna platform. The component is also provided with a sighting surface to permit alinement parallel to the lubber line of the aircraft.

RESOLVER SYNCHRO.—Within the resolver synchro, the pitch and roll signals from the gyro are combined by induction in a rotor coil that is geared to the rotary head of the scanner, with which it rotates in a one to one ratio. Because of this mechanical coupling, the resultant resolver-synchro signal used in tilt stabilization experiences the proper variation throughout the scanning cycle to maintain the selected reflector tilt angle with respect to the aircraft for every azimuth position of the scanner. The net effect of tilt stabilization is to maintain a fixed predetermined angle of tilt between the reflector and the true horizontal, while the aircraft experiences the usual variations of normal flight.

The pitch and roll signals developed by the gyroscope unit are fed into the separate stator windings of the resolver synchro. These stator windings are mounted at right angles to each other. The orientation of the rotor is adjusted so that when the reflector is dead ahead, pitching of the aircraft causes maximum pitch signal to be induced in the rotor of the resolver synchro, and when the reflector is directed to extreme port or starboard, rolling of the aircraft causes maximum roll signal to be induced in the rotor of the resolver synchro. At intermediate positions of the reflector, combinations of both pitch and roll signals are induced in the rotor. One of these voltages may be either aiding or opposing the other, according to the attitude of the aircraft, with respect to the azimuth directivity of the reflector.

SERVO AMPLIFIER.—The SERVO AMPLIFIER amplifies the error signal resulting from the combining of the manual-tilt synchro signal and the resolver-synchro signal, and controls the voltage to the tilt-torque motor, which corrects the tilt angle of the scanner.

The servo amplifier uses two resistance-coupled amplifiers, followed by a phase-discriminator stage and a pair of d-c amplifiers working into a saturable-reactor assembly. The output of the phase-discriminator stage, amplified by the d-c amplifiers, controls the a-c voltage that drives the tilt-torque motor. The direction of drive, which determines direction of tilt, is governed by the phase relationship between the exciting voltage and the control voltage of this motor; this relationship is, in turn, dependent upon the phase relationship between the motor-exciting voltage and the error voltage at the input of the servo amplifier.

The scanner is maintained at an angle of tilt with respect to the horizon. This angle is determined by the setting of the TILT control. This condition is achieved by combining the signal from the manual-tilt synchro with the one from the resolver synchro at the input of the servo amplifier. The signal from the manual-tilt synchro is applied to one side of the ratio potentiometer, *R*1, and the signal from the resolver synchro is applied, through resistor *R*2, to the opposite terminal of the same potentiometer, whence they are combined in the servo amplifier. (See fig. 8–9.) These two signals are 180 degrees out of phase, and when of equal amplitude they cancel each other. When this cancellation is complete, there is no control voltage for the tilt-torque motor, and consequently no tilt action is applied to the scanner.

When the aircraft rolls, the resolver-synchro signal is altered in amplitude, provided the scanner is not directed dead ahead or aft (at zero degree or 180 degrees azimuth). When the aircraft pitches, the resolver-synchro signal is altered in amplitude, provided the scanner is not directed at 90 or 270 degrees relative azimuth. As a result, the resolver-synchro signal and the manual-tilt-synchro signal no longer cancel, and an error signal is developed at the input of the servo amplifier. After amplification, this error signal is applied to a phase discriminator, which determines from the phase of the signal whether the reflector must be tilted up or down in order to correct for the roll, pitch, or combination of the two that has just caused the error signal. The output of the phase discriminator drives the proper tilt-servo control tube, which then applies the control voltage to the tilt-torque motor by means of a saturable reactor. As the tilt-torque motor corrects the scanner tilt angle, the rotor of the followup synchro, which is coupled to the tilt-torque motor, is turned in such a direction that the manual-tilt synchro signal changes in amplitude to compensate for the change in the resolver-synchro signal. When these two signals cancel again, the necessary correction to the reflector tilt angle has been accomplished, and the tilt-torque motor stops.

The ratio potentiometer, *R*1, balances the signals from the manual-tilt synchro and the resolver synchro to compensate for possible deviation from the true horizontal of the mounting plane of the gyroscope unit.

This servo amplifier functions similarly to the one discussed under Azimuth Stabilization in this chapter, in that the error voltage controls the saturable reactor which provides the variable voltage for the control winding of the tilt-torque motor. (See fig. 8–9.)

In this tilt-stabilization system one winding of the TILT-TORQUE MOTOR is continuously excited by 115 volts a.c. The other winding, used to control the starting and operation of the motor, is supplied from the saturable reactors with a-c voltage developed whenever an error signal is amplified by the servo amplifier. The motor torque is obtained by providing approximately 90 degrees phase difference between the control voltage and the excitation voltage, by means of capacitor *C*10 which is shunted across the control winding. The speed of the motor is dependent upon the amplitude of the control voltage, which in turn is dependent upon the action of the TILT control, tilt stabilization, and damping.

MANUAL-TILT CONTROL.—The scanner tilt angle is controlled by the tilt synchro in the radar set control and is adjusted by means of the TILT control. The tilt synchro is connected with the followup synchro in the tilt-servo unit. Supplied with a steady a-c voltage, this synchro circuit develops the manual-tilt-synchro signal in its rotor to control the tilt angle of the beam with respect to the angle of tilt stabilization (approximate horizon). The signal from the manual-tilt synchro is fed to the input of the servo amplifier, where it is combined with the resolver-synchro signal. Control of the scanner tilt angle is therefore dependent upon the resultant sum of these two voltages, and is accomplished through the system used for tilt stabilization, as just described. If the resolver-synchro signal is switched off by means of the STAB–UNSTAB switch *S*2 on the servo amplifier, the beam-tilt angle is then established only with respect to the aircraft.

FOLLOWUP SYNCHRO.—The followup synchro and the manual-tilt synchro develop the manual-tilt-synchro signal. The followup synchro is located in the tilt-servo unit of the antenna assembly. The rotor of this synchro is mechanically coupled to the tilt mechanism so that the rotor movement occurs in synchronism with the tilting of the scanner. The stator windings of the followup synchro are connected to the stator windings of the manual-tilt synchro in the radar control unit. The rotor of the followup synchro is supplied with 26-volt, 400-cycle, single-phase voltage. Voltages of unequal amplitude are induced in the stator windings (fig. 8–9). These voltages are reproduced in the stator windings of the manual-tilt synchro in the radar set control. When the rotor position of the manual-tilt synchro is such that its signal cancels the resolver-synchro signal and produces a null at the input of the servo amplifier, no control voltage is developed for the tilt-torque motor and the scanner tilt angle is not changed. When the rotor is displaced from this position in either direction by resetting the TILT control for a new beam tilt, one signal becomes larger than the other, cancellation is no longer complete, and therefore an error signal is developed which causes the servo amplifier to develop a control voltage for the tilt-torque motor. As the motor drives the scanner to the required tilt angle, a mechanical coupling from the motor turns the rotor of the followup synchro. When the manual-tilt synchro again cancels the resolver-synchro signal at the input of the servo amplifier, the tilt-torque motor stops. The beam-tilt angle thus obtained is determined with respect to the horizontal.

The function of the DAMPING-SIGNAL GENERATOR that is used with this system of tilt stabilization has already been explained in this chapter in connection with Antihunt Devices. This information will not be repeated. It is of interest to note that the hunting of the scanner in this equipment is held to a maximum of plus or minus one-half degree of tilt about the angle of stabilization. This, as you can easily understand, is a very close tolerance of error and thus, hunting is adequately eliminated.

QUIZ

1. Which of the following statements best describes a servomechanism system?
   a. A system for transmission of angular information
   b. A power amplifier to supply a split phase induction motor
   c. An error indicating system to show the displacement of one shaft from another
   d. A system that transmits angular information with an amplification of power

2. A block diagram of servo controller components would always include a/an
   a. modulating device
   b. vacuum tube device
   c. phase sensitive device
   d. electromechanical generating device

3. In a servomechanism system, the controller is the
   a. synchro generator
   b. error indicator
   c. servo amplifier
   d. feedback device

4. The error detector is usually a
   a. differential generator
   b. synchro control transformer
   c. synchro generator
   d. magnesyn

5. The main disadvantage of using a multiple-speed error detector is
   a. the need of an additional synchro system and switching circuit
   b. that they tend to be inaccurate
   c. the possibility of locking-in on an incorrect position
   d. that the mechanical complexity is impractical

6. The component of the servo amplifier in figure 8–4 that transforms the a-c error signal into a d-c signal is the
   a. sense detector
   b. a-c amplifier
   c. d-c amplifier
   d. modifying network

7. When a two-phase servomotor is used to position a load, the voltages supplying the two field windings should be
   a. 180 degrees out of phase
   b. in any phase relationship
   c. in phase
   d. 90 degrees out of phase

8. The plates of the diodes (fig. 8–5) are fed
   a. push-pull by the reference voltage and push-pull by the error voltage
   b. push-push by the reference voltage and push-pull by the error voltage
   c. push-push by the reference voltage and push-push by the error voltage
   d. push-pull by the reference voltage and push-push by the error voltage

9. With no error voltage applied to *T*2 (fig. 8–6), the exciter field flux would be
   a. in phase with the reference voltage
   b. maximum in intensity
   c. minimum in intensity
   d. varying in intensity

10. With reference to figure 8–7, when the error voltage at terminal 3 of *T*1 is of the same polarity as the reference voltage at terminal 3 of *T*4,
   a. *V*4–*A* or *V*3–*B* will conduct and the output at terminal 6 of *T*2 will be in phase with the reference voltage
   b. *V*4–*B* or *V*3–*A* will conduct and the output at terminal 6 of *T*2 will be 180 degrees out of phase with the reference voltage
   c. *V*4–*A* or *V*3–*B* will conduct and the output at terminal 6 of *T*2 will be 180 degrees out of phase with the reference voltage
   d. *V*4–*B* or *V*3–*A* will conduct and the output at terminal 6 of *T*2 will be in phase with the reference voltage

11. The purpose of feedback as used in a servomechanism (fig. 8–8) is to
    a. cancel hunting
    b. position the control transformer for zero error signal
    c. turn control transformer stators
    d. cancel error signals

12. To prevent overshoot in a servomechanism, without sacrificing accuracy and speed of response, it is a good idea to
    a. apply some additional friction to the output shaft
    b. reduce the gain of the servo amplifier
    c. utilize a damping signal generator
    d. utilize an inertia damper on the output shaft

13. Inertia dampers are used on
    a. synchro transmitters
    b. synchro receivers
    c. control transformers
    d. differential generators

14. Which one of the following antihunt devices utilizes a resonant circuit in its operation?
    a. The limiter type
    b. The inertia damper type
    c. The damping signal
    d. The derivative type

15. The balance control in a servomechanism is adjusted
    a. for the fastest response without hunting
    b. for a maximum output when no error signal is applied
    c. to compensate for variations in line voltage
    d. for zero output when no error signal is applied

16. What is the purpose of using a sense antenna in addition to the loop antenna for loop control (fig. 8–14)?
    a. To eliminate one of the loop antenna nulls
    b. So that radio signals will not be affected by repositioning of the loop
    c. To prevent fading due to night effect
    d. To utilize both the electromagnetic and the electrostatic fields to operate the synchro system

17. In a null condition, the signal present at the secondary of *T*3 (figs. 8–15 and 8–16) consists of
    a. an RF signal from the sense antenna only
    b. a 100-cycle signal output from the balanced modulator
    c. combined RF signals from the loop and sense antennae
    d. an RF signal from the loop antenna only

18. Which of the following statements is correct concerning the signals on the grids of the thyratrons?
    a. Signals are in phase with the reference voltage.
    b. Signals are out of phase with the reference voltage.
    c. Signal phase depends upon whether the transmitter is to the right or left of the aircraft.
    d. Signals are of the correct phase to fire both thyratrons simultaneously.

19. The purpose of azimuth stabilization is to
    a. facilitate determination of relative bearing of targets
    b. facilitate low altitude bombing
    c. allow targets to change position as your heading is changed
    d. maintain the top of the indicator at true or magnetic north

20. The purpose of the zero set synchro (fig. 8–17) is to
    a. electrically zero the synchro system
    b. aline the sweep on the indicator with the antenna on relative bearing operation
    c. correct for compass error on true bearing operation
    d. aline the sweep on the indicator with the antenna on true bearing operation

21. What would be the output from the true bearing amplifier circuit (fig. 8–19) if *R*1 (sense pot) was moved all the way down?
    a. Zero
    b. Constant phase and amplitude
    c. Constant phase with varying amplitude
    d. Varying phase with constant amplitude

22. What is the purpose of tilt stabilization?
    a. To keep the antenna pointing dead ahead of the aircraft
    b. To prevent the antenna from following changes in flight angle
    c. To allow the antenna to follow changes in flight angle
    d. To prevent loss of targets when maneuvering by stabilizing the aircraft with respect to the earth's surface

23. The antenna hunts up and down continuously. (See fig. 8–9.) To correct this condition adjust
   a. *R*1 up
   b. *R*1 down
   c. *S*1 to the right
   d. *S*1 to the left

24. The component of the tilt stabilization system (fig. 8–9), which interprets the amount of correction to be applied in order to correct the antenna's position in a vertical plane to coincide with its position in the azimuth plane, is the
   a. resolver synchro
   b. followup synchro
   c. manual tilt synchro
   d. servo amplifier

CHAPTER 9

MAGNETIC AMPLIFIERS

Electronics personnel usually associate amplification of voltage or power with vacuum and gas tubes. However, several devices accomplish amplification without the use of tubes. One of these, the MAGNETIC AMPLIFIER, which has been in use here and abroad for several years, will be discussed in this chapter; a second device, the TRANSISTOR, is a recent American invention that shows great promise for future use and will be discussed in chapter 10.

The term amplification, in general, refers to the process of increasing the amplitude of voltage, current, or power. The term AMPLIFICATION FACTOR is the ratio of output to input. The input is a signal that controls the amount of available power delivered to the output. Thus, tubes are really valves that permit a controlled amount of power to be delivered to a load. A difference in high-vacuum tubes and gas tubes is that the signal on the grid of a vacuum tube always has control of the output signal—that is, the valve can open and close partially, whereas in gas tubes the grid loses control when the gas ionizes, so that the valve is either completely open or closed. For this reason thyratrons frequently are supplied with a-c plate voltage—the valve shuts off automatically when the plate voltage cycles to zero. It will be seen that the magnetic amplifier is comparable in action to a thyratron.

Magnetic amplifiers are not new devices by any means, having been reported in use over fifty years ago. Their popularity is just now beginning to be widespread and their capabilities realized by electronics engineers. A magnetic amplifier is essentially a device which controls the a-c reactance of a coil by utilizing a d-c signal to modify the permeability of the magnetic material upon which the coil is wound.

In the early stages of development this basic idea was incorporated into devices, usually designated as saturable reactors, and used to control large electrical loads such as theatrical lighting and electric furnaces. However, the use of the SATURABLE REACTOR, the heart of the magnetic amplifier, was limited in its ability to control the load until the recent development of improved magnetic materials and efficient metallic rectifiers. The utilization of these improved materials resulted in the use of the saturable reactor in more elaborate circuits and led to the distinguishing term MAGNETIC AMPLIFIER. Many different trade name devices have been used to identify the more elaborate saturable reactor circuits. Nevertheless, in this chapter the term magnetic amplifier will be used to identify all amplifying systems which utilize the saturation of a magnetic core for the purpose of control. It is felt that this is the most popular of all terms used at the present time.

The advantages of the magnetic amplifier are principally based on the fact that it is a completely static device. With the exception of the rectifiers used, its mechanical construction is comparable to that of an iron-core transformer. There are no contacts, moving parts, filaments, or other features which account for most of the failures associated with other types of amplifiers (except for those using transistors). The need for frequent inspection and maintenance is cut to a minimum. The life of the magnetic amplifier is more or less indefinite, and it is especially suited for aircraft installation where there are adverse operating conditions such as vibration and shock.

In order to understand the theory of magnetic amplifiers, it is necessary that you possess a knowledge of magnetism and magnetic circuits. This information may be found in the Navy Training Course *Basic Electricity,* NavPers 10086, chapter 7.

BASIC PRINCIPLES OF OPERATION

Basically, a magnetic amplifier is nothing more than a controlled variable inductance in series with an a-c power source and a load. It may be recalled that any coil possesses a property of electrical inertia called inductance. The inductance of a coil may be increased by providing it with an iron-core magnetic circuit. The inductance of coils with magnetic cores may be determined with fair accuracy by the following formula:

$$L=\frac{1.256N^2A\mu10^{-8}}{l}$$

where

L = Inductance in henrys

N = Number of turns

A = Area of core in square centimeters

μ = Permeability of magnetic material

l = Length of core in centimeters

Permeability is a measure of the relative ease with which flux flows in a material. The permeability of air is one (1). The permeability of the core of a coil can be increased if the air is replaced by some type of ferromagnetic material such as iron or steel. By observation of the preceding formula, it can be clearly seen that if the permeability (μ) is increased by one thousand (1,000) when the iron core is inserted, that the inductance will be increased one thousand (1,000) times over the inductance of the same coil with an air core.

Figure 9–1 shows a simple magnetic amplifier where the iron core is mechanically moved in and out of the coil to vary the inductance and hence the inductive reactance of the coil. When the iron core is completely within the

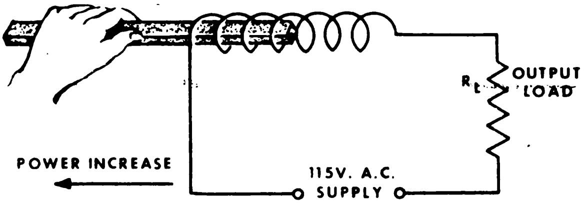


Figure 9–1.—Varying coil inductance mechanically.

coil, the inductance of the coil is maximum. Therefore, the voltage drop across the inductance is maximum, and so reduces the voltage across the load resistor R_L to a low value. However, removing the core from the coil lowers the inductance of the coil (hence the impedance of the circuit), decreasing the voltage across the inductance and increasing the voltage across the load to nearly 115 volts. In effect, this is the operating principle of a magnetic amplifier. Since it requires relatively few watts of power to move the core within the coil, which in turn can control perhaps several horsepower, the device is an AMPLIFIER.

Although the mechanical arrangement is effective, and some types are still in use, the present day method of controlling the permeability of the coil core is obtained by saturating the core with a relatively small amount of d-c or properly phased a-c voltage. An unsaturated core provides a relatively high impedance to the circuit. A saturated core, however, acts similar to an air core, offering practically no impedance except for the low d-c resistance of the coil.

By briefly reviewing some of the facts concerning saturation of a coil, it will be remembered that when current flows through a coil there is set up about the coil a magnetomotive force, the intensity of which depends upon the amount of current flowing through the coil. This force in a magnetic circuit may be compared to the voltage of an ordinary electric circuit. The magnetomotive force sets up a flux about the coil, which is comparable to the current of an ordinary circuit and has a density which depends on the reluctance of the core of the coil. The core reluctance, which may be compared to the resistance of an electric circuit, has a value which depends on the material forming the core. The reluctance of an air core remains constant regardless of current. This results in an increase of flux density which is proportional to the increase in both current and magnetomotive force. When a magnetic substance makes up the core of a coil, however, the reluctance is no longer constant, regardless of current. Instead, as a current begins to flow, the reluctance is very low and the flux is very high compared to that existing in an air-core coil under similar conditions. With an increase in current, the reluctance gradually increases and the rate-of-flux increase is reduced. After the current reaches a certain value, which depends on the core material used, the reluctance increases very rapidly until its value approaches that of air. In this region, around point *b* in figure 9–2, any further increase in current will produce no appreciable increase in flux. This condition is known as SATURATION.

As previously stated, a saturated core acts similar to an air core for it offers practically no impedance except for the low d-c resistance of the coil. This stems from the fact that the inductance of a coil is proportional to its incremental permeability, which is the ratio of a small change of flux density, ΔB, over a correspondingly small change in magnetic field intensity or ampere turns, ΔH. (Incremental $\mu = \Delta B / \Delta H$.)

This can be graphically shown by considering a typical magnetization curve, such as that shown in figure 9–2. The flux density, B, is shown in kilogausses and the magnetic field intensity, H, is shown in oersteds which are proportional to ampere turns. The inductance of the coil is proportional to the rate of change of flux density, B, with respect to magnetic field intensity (ampere turns), H. It will be seen that in the region of the curve around point *a*, a small change of ampere turns, ΔH, will produce a large change of magnetic flux density, ΔB. This

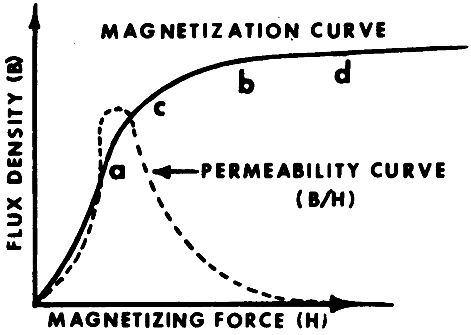


Figure 9–2.—Magnetization and permeability curves.

represents the unsaturated condition of the coil, and therefore its inductance is large.

In the region of the curve around point *b*, the same small change of ampere turns, *H*, will now produce only a very small change in flux density, *B*. This represents the saturated condition of the coil and its inductance in this region is very low.

An application of the basic saturable reactor may be used in a special type of filter choke that is often used in power supplies where the current drain varies considerably. They are called swinging chokes and are designed to operate on the portion of the curve between points *a* and *b*. For low values of current, their inductance values are quite high. For high values of current their inductance values are low since the core has become saturated. The air gap in the swinging choke is much smaller than in the ordinary filter choke and, consequently, the total reluctance or opposition to the magnetic lines of force through the core is small. This permits the choke to become saturated for high values of current. This type of choke has special applications in circuits where the direct-current variations are quite large, since it has been found that they give better direct-voltage regulation with varying amounts of direct current. That is, the voltage remains nearly constant even though the amount of current varies over a wide range.

The magnetic amplifier works on the principle that varying the average magnetic field intensity will vary the inductance. Referring again to figure 9–2, the curve from the origin to the region below point *c* is quite steep but approximately straight. The inductance in this region is therefore high but constant, that is, given changes of ampere turns will produce high but constant changes in flux density. Operating the coil in this region will therefore produce little or no change in inductance for changes in ampere turns and no amplifier action will result.

The region of saturation around point *b* is likewise very flat, so that the inductance around this region is very small and changes very little with changes in ampere turns. Operation of the coil in this region will likewise be unsuitable for amplifier action.

In the curved or nonlinear portion of the curve around point *c*, the inductance changes quite rapidly with small changes of ampere turns; that is, the rate of the rate of change of flux density with respect to ampere turns is large. This region is the desired operating region of the coil for magnetic amplifier action because small changes of average ampere turns produce large and significant changes of inductance, therefore producing high gain.

It is possible to derive the permeability curve for the magnetization curve shown in figure 9–2 since the ratio B/H is the slope of the magnetization curve. This permeability curve is represented by the broken line in the figure. It clearly shows that the permeability is low for small and very large values of magnetizing force and is rather high for intermediate values of magnetizing force. The inductance of a coil can be varied by altering the magnetizing force. As illustrated in figure 9–2, as the MAGNETIZING FORCE IS ALTERED the permeability is varied, hence the inductance is also varied.

Removing the iron core physically, as shown in figure 9–1, does not actually control the effective permeability of the magnetic material since the core is actually replaced by air. In most modern applications, the saturation method is used. This not only simplifies the installation but permits a much higher speed of response, eliminates mechanical moving parts, and considerably widens the range of application.

THE BASIC CIRCUIT

The basic circuit for controlling the load power by means of a control winding is illustrated in figure 9–3. Both windings are on a common core; the primary is usually referred to as the control winding.

The primary winding is supplied with a d-c source for control purposes. A potentiometer is connected in series with this circuit and is used to control the amount of current flow. This current flow will establish a unidirectional flux (gausses) in the core at a level depending upon the amount of d-c magnetizing force (oersteds). (Oersted×2.021=ampere turn/inch; gauss×6.452=line/square inch.) The second winding is connected in series with an a-c source and the load, which is shown as a lamp.

As can be seen in figure 9–2, changes of current through the control winding corresponding to variations in magnetizing force (H) between the origin and point *a* will produce little or no change in inductance due to the nearly straight characteristics of the curve. In order for these changes in current to produce changes in inductance, they must take place on the curved portion, or **KNEE** of the curve. This means that sufficient current must flow through the control winding to place the operating point at the knee, such as at point *c*. This is very similar to setting the operating point of a vacuum tube by means of bias. With the coil operating on the knee of the magnetization curve, a small increase of current through the control winding will lower the inductance of the coil, or permeability of the core, because the magnetization curve's direction or slope is moving toward a horizontal direction. As a result, the a-c reactance of the load wind-

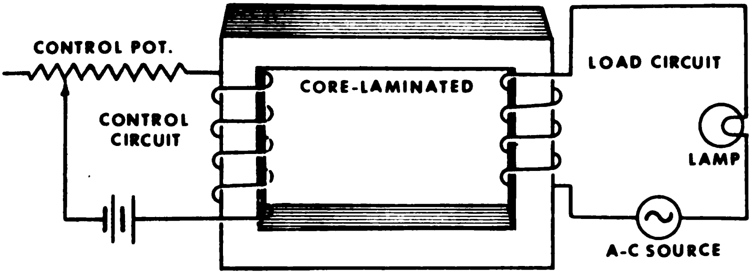


Figure 9–3.—A simple magnetic amplifier.

ing is decreased. The decreased impedance of the load circuit will then allow a larger current to flow, thereby increasing the power developed across the load. With the core completely saturated, the load winding reactance drops to nearly zero, leaving only the resistance of the copper wire and load to impede the flow of current. Under this condition, the maximum supply voltage will be applied to the load, thus the lamp will be brightest. Conversely, decreasing the current through the control winding will increase the reactance of the load coil. In the region between the origin and point *a* on the curve in figure 9–2, the core permeability, hence coil inductance, is maximum, resulting in minimum load current and, consequently, minimum power. Since the reactance of the load coil winding has been changed by varying the degree of saturation, the device is referred to as a **SATURABLE REACTOR.**

Although the explanation just given shows the basic way of controlling the core permeability, the circuit of figure 9–3 is a very inefficient magnetic amplifier because of transformer action between the two windings. The alternating flux from the secondary will induce a voltage into the control winding. If the control winding has a large number of turns the voltage may become excessive and may even break down the insulation. Another reason why this form of magnetic amplifier is seldom used is due to the fact that the control winding would act as a short-circuited secondary. This would dissipate a considerable amount of a-c energy that is applied to the output or controlled winding. It is impossible to prevent this loss by inserting an isolating impedance, in the form of an inductance, in series with the control winding. Another method of preventing this loss is by utilizing a THREE-LEGGED core. This method will be described in detail.

The transformer action previously mentioned would also cause an oscillating flux in the core and the usual *B–H* hysteresis loop of the core material would be traced out during one full cycle of the a-c source voltage as in transformer operation. In order for it to function as an amplifier, sufficient d-c magnetizing force (oersteds) will be required to counteract the a-c flux, and an additional amount of d-c magnetizing force will be needed to set the operating point. The characteristics of the circuit shown in figure 9–3 are low gain, but a linear control action.

A much more satisfactory arrangement for a saturable reactor is shown in figure 9–4. This figure shows a THREE-LEGGED CORE having an a-c winding on each outer

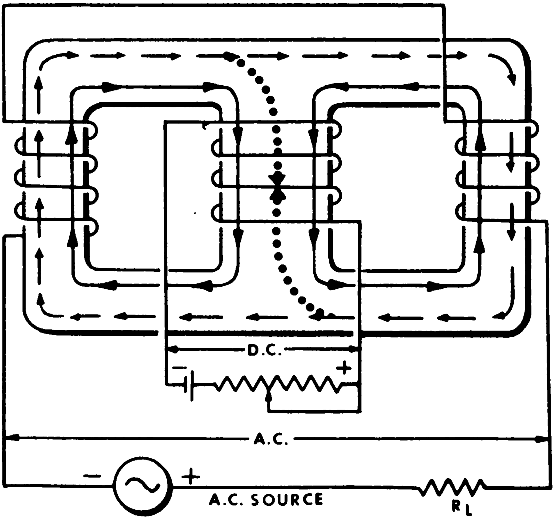


Figure 9–4.—A magnetic amplifier with three-legged core.

leg and the d-c control winding on the center leg. In a saturable reactor of this type, any alternating flux through the middle leg that is produced by one of the a-c coils is balanced by the flux produced by the other a-c coil. This is true only when both a-c coils have an equal number of turns and are connected either in series or parallel such that the flux produced by the two coils oppose each other through the center leg. This is shown by the dotted lines in figure 9–4.

Since these forces are opposite and equal, they do not pass through the center leg but join in a common path through the outer legs of the core. This is shown by the broken lines in figure 9–4. The figure also shows that the current through the d-c winding on the center leg of the core will produce a magnetic flux as indicated by the arrows on the solid lines. The flux lines caused by this direct current flow in two paths, as shown, and magnetize the entire core. Thus, it has been shown that although the d-c coil is used to magnetize the cores of the a-c coils, no a-c voltage is induced into the control coil when a three-legged reactor is used.

During operation, variations of current in the control coil will vary the magnetization of the core. As has been shown previously, when the magnetizing force is altered, the permeability of the core is varied and, in turn, the inductance of the a-c coils is also varied. In other respects the three-legged magnetic amplifier functions similar to the basic circuit that is illustrated in figure 9–3.

MAGNETIC AMPLIFIER WITH HALF-WAVE RECTIFIER.—Although figure 9–3 demonstrates the basic principle used to control core saturation, this type of magnetic amplifier is relatively inefficient. This is true because the load supply voltage, which is a.c., will create a sinusoidally varying flux which will flow through the core in one direction one-half of the cycle and in the opposite direction during the other half. Thus, half of the time the flux created by the load current will impede the d-c control winding flux and the other half of the time it will

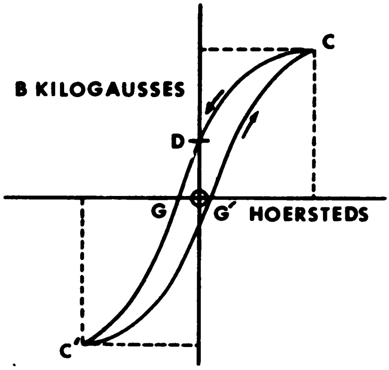


Figure 9–5.—Hysteresis loop.

aid this flux. This will shift the operating point of the amplifier from the origin (see fig. 9–2) to a point of the curve whose abscissa (horizontal coordinate) corresponds to the peak value of ampere turns, back to the origin, and along a similar curve in the negative direction. In other words, the operating point of the amplifier will trace out a hysteresis curve such as is shown in figure 9–5. Because the amplifier cannot operate in this manner, the oscillating flux must be BUCKED or counteracted and then a directional flux must be added to partially saturate the core or place the operating point on the desired portion of the magnetization curve.

In order to produce sufficient flux to balance out the oscillating flux caused by the a.c. in the load winding, the ampere turns applied to the control winding would have to equal the ampere turns of the load winding. Then sufficient extra ampere turns would have to be applied to the control winding to set the operating point at the desired place on the magnetization curve. It is obvious then that this would be a very inefficient amplifier because the ampere turns required by the control loop would be greater than the ampere turns in the load circuit.

A more efficient INTERNAL feedback magnetic amplifier circuit is shown in figure 9–6. A half-wave rectifier has

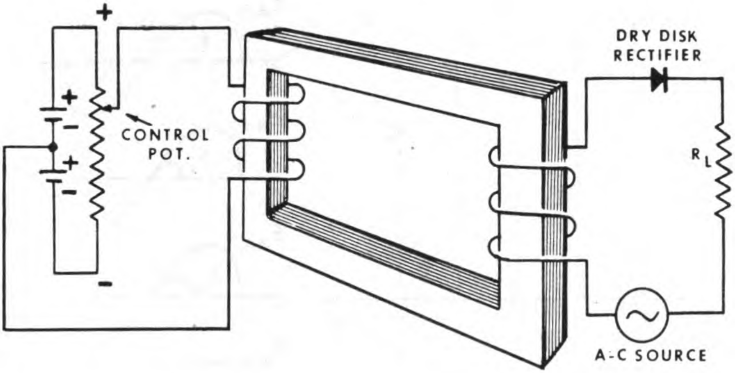


Figure 9–6.—Magnetic amplifier with half-wave rectifier.

been inserted in the load circuit which allows current to pass in one direction only. This unidirectional load current eliminates the oscillating flux caused by the a.c. in the load circuit. Thus, the load current assists the control winding current in saturating the core. A more efficient amplifier is provided since less control power is needed for a given amount of control. In this figure the control circuit is so arranged that the magnetic flux resulting from the control current may either aid or oppose that resulting from the load current.

A brief explanation of the self-saturating half-wave magnetic amplifier is possible by using the magnetization curve shown in figure 9–7. This curve has been drawn with the effect of hysteresis neglected. Positive *a* on this curve represents the condition of the control flux when the potentiometer in figure 9–6 is set at the positive end. Under this condition the control current produces a flux which is in opposition to the flux created by the load current. If the amplitude of the alternating current is such that it varies the core flux along the magnetization curve from point *a* to point *d* the impedance of the secondary remains large and the voltage drop across the load will be approximately zero. The relation of the applied voltage and load current for this type of operation is repre-

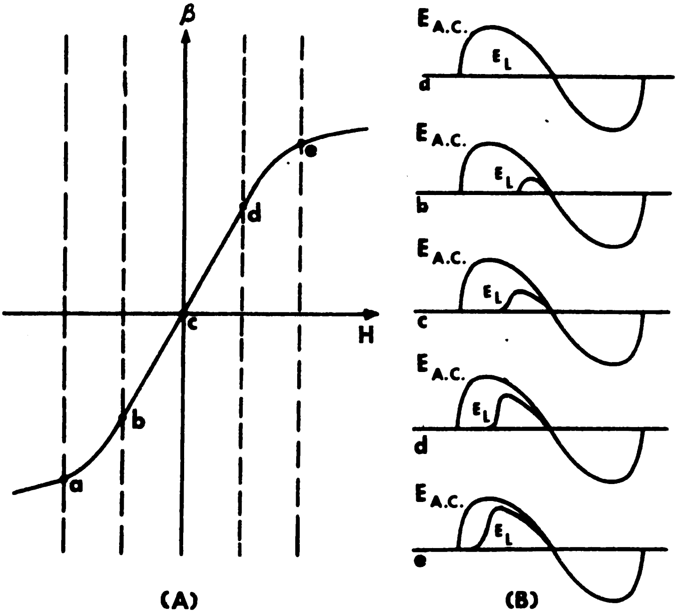


Figure 9–7.—Operation of a self-saturating magnetic amplifier.

sented by *a* in figure 9–7 (B). When the potentiometer is moved toward the negative end to the position where current still flows in the same direction but of less amplitude so that the control flux is at point *b* on the curve, the rectified alternating current will now vary the core flux from point *b* to point *e*. Under these conditions partial saturation of the core is achieved resulting in a low voltage waveshape as shown at *b* in figure 9–7 (B). As the potentiometer is moved further toward the negative end, the initial flux density in the core (no load winding flux) is nearer saturation so that the inductance of the controlled winding is reduced earlier in the half cycle and larger values of load current flow. This is shown by *c*, *d*, and *e* in figure 9–7 (B).

EFFECT OF HYSTERESIS ON OPERATION.—Up to this point, in order to simplify the introduction to magnetic amplifiers, the effect of hysteresis has been neglected. In

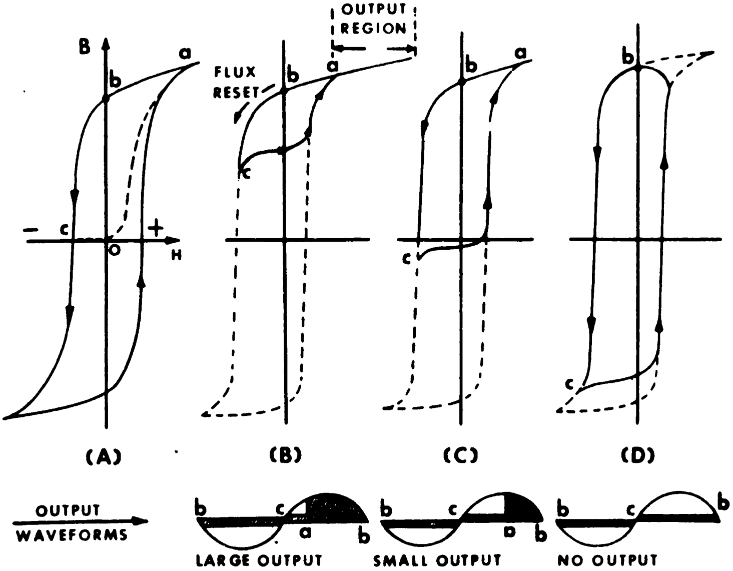


Figure 9–8.—Operation of a magnetic amplifier showing effects of hysteresis.

order to fully understand the operating characteristics of magnetic amplifiers, it is important to know how these characteristics are affected by the shape of the hysteresis loop.

The curve in figure 9–8 (A) represents a typical rectangular-shaped *B–H* curve; the curves in (B), (C), and (D) show the operation for various amounts of control action.

Further use of figure 9–6 will be made in explaining the effects of hysteresis. This circuit actually has two distinct periods of operation. The period during which the rectifier conducts is known as the "operating period"; the other half cycle of the a.c. is known as the "control period."

In the explanation of the effect of the hysteresis loop of figure 9–8 (A), assume that control current is zero and that the load winding current is of such amplitude that it will saturate the core during peak conductance. During the first operating period, conduction through the load winding will cause the core flux to build up along the dotted line from the point of origin, *o*, to the point of saturation, *a*. At the end of this period the flux level in the coil returns to point *b* rather than to zero because of the retentivity of the core. This is called the remanence point in magnetics and leaves the core in a magnetic state, similar to a permanent magnet. Thus, with no control current the core flux will remain at point *b* and during each positive half cycle the core will saturate immediately. Under this condition the load current will be maximum for no control is exercised.

Figure 9–8 (B) shows the action when a small control current is added. The operation during the first half cycle is similar to that just described in connection with figure 9–8 (A), but during the second half cycle (control period), the flux will reset from point *b* to point *c*. This occurs because the direction of the control flux in the core is opposite to that of the operating flux.

This positioning of the residual flux is called "resetting." During the next half cycle the flux starts increasing from point *c* in the figure and a small period of this half cycle is used before the flux reaches saturation at point *a*. At this point the core is saturated and the rectifier is in maximum conduction for the remaining part of this half cycle. At the end of the operating period the flux returns to point *b* where the control period starts and resets the flux to point *c* where the cycle starts repeating itself. The output for this condition is shown as the shaded portion of the waveform of figure 9–8 (B). Figure 9–8 (C) and (D) shows the operation for increasing amounts of control action. The operation in each case is basically the same as for figure 9–8 (B).

It may be seen that the action just described is similar to that of a thyratron. The control signal sets the point of amplitude at which the source voltage causes saturation of the core and reduces the impedance in series with the load winding to a very low value. In the thyratron the control signal on the grid sets the point of amplitude at which the source voltage causes the tube to ionize, which results in a decreased impedance. The term FIRING, as it relates to the thyratron, is frequently used in connection with magnetic amplifiers interchangeably with the term SATURATION.

USING BIAS FOR FLUX RESET.—When the core material possesses hysteresis properties that produce a rectangular-shaped *B–H* curve, it may be necessary to bias the core in order to retain control. This is accomplished by allowing a bias current to either flow through the control winding or through a separate bias winding. This bias current provides a means of resetting the flux to the initial operating point during the control period. Thus, no control current is necessary to reset the flux. Figure 9–9 shows a separate bias winding wound on the center leg of a three-legged core.

The core is usually biased so that with zero control current the load winding flux saturates the core midway in the operation cycle. This bias setting corresponds roughly to point *c* of figure 9–8 (C), and enables the control current to either advance or delay the point of saturation. In other words, if the polarity of the control current is such that the flux produced by it adds to the bias flux, the point of saturation will be delayed. A con-

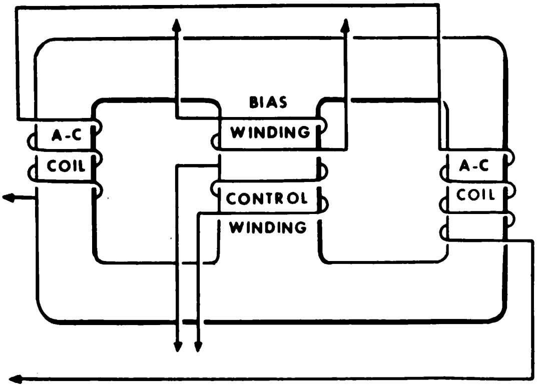


Figure 9–9.—Bias winding.

trol current of opposite polarity will produce a flux that opposes the bias flux and this will advance the point of saturation.

The bias may be either alternating or direct current. In most cases the polarity of d-c bias is such that it opposes the LOAD current flux. The magnitude of the bias might be such that it would reset the flux to a point between *a* and *c* of figure 9–2. There are some applications where the opposite is true; in these applications the purpose of the bias is to provide an initial d-c saturation in the core in order to obtain greater amplification of weak signals.

MAGNETIC AMPLIFIER WITH FULL-WAVE RECTIFIER.—The magnetic amplifier previously discussed and shown in figure 9–6 produces a pulsating or half-wave output. In most applications of magnetic amplifiers full-wave operation is more desirable for it is more efficient since the load is energized during both halves of the a-c cycle. Full-wave operation may be obtained by using a pair of simple half-wave units. A typical circuit arrangement is shown in (A) of figure 9–10.

The load current is controlled by means of two control windings connected in series. Each amplifier unit contains a rectifier so that load current flows alternately in the two amplifiers. As one conducts, the core of the series coil in the other is being set by the action of the control current. And with each amplifier conducting approximately one half cycle of load current, the output variations are full-wave in nature.

Output waveforms for two different control currents are shown in (C) and (D) of figure 9–10. The waveform in (C) is produced when the control current biases the core near the point of saturation, such as point *e* in figure 9–7 (A). In this condition, saturation of the magnetic cores of the amplifiers is reached early in each half cycle, and the resulting output current variations are almost sinusoidal in shape.

When the current in the control winding biases the

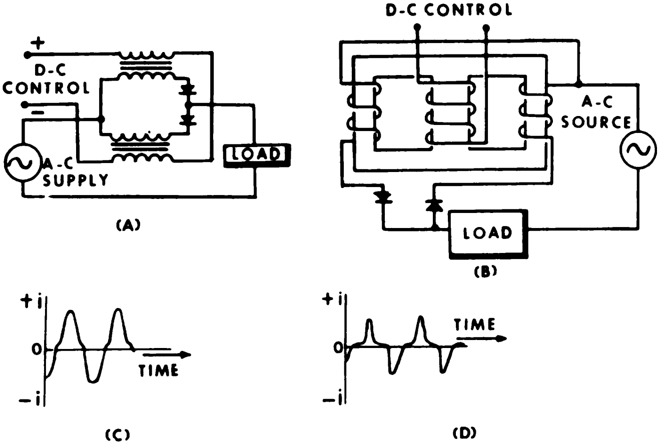


Figure 9–10.—Full-wave magnetic amplifier and output waveforms.

core to point *c* of the curve in figure 9–7 (A), the output resembles the waveform of figure 9–10 (D). The average value of the load current is now considerably less than when the control current biased the core to near saturation. Thus, as in the half-wave circuit, the output power developed in the load can be varied by controlling the load current. The output that is developed is an alternating current and is in phase with the source voltage.

Figure 9–10 (B) shows a magnetic amplifier that functions similar to the one shown in (A). The major difference is that the two cores shown in (A) have been combined to form a three-legged core. Another possible arrangement, which prevents interaction and also conserves space, is to use two separate circular cores that are stacked coaxially. Each core contains an individual a-c winding. The cores are then stacked so that the a-c windings are wound in opposite directions. A common control winding is wound around both cores after stacking. This arrangement prevents a.c. from being coupled into the control winding.

FEEDBACK IN MAGNETIC AMPLIFIERS.—It is possible to change some of the characteristics of a magnetic amplifier by the use of an external feedback circuit. (NOTE: The action of rectifiers in series with the load current produces internal feedback.) External feedback may be provided by winding an additional coil on the center leg of the core as illustrated in figure 9–11.

The circuit in (A) of figure 9–11 is similar to the one shown in figure 9–10 (B) except that a full-wave bridge rectifier is used to provide a d-c output to the load. Dry disk rectifiers are usually employed for this purpose. The circuit in (B) of figure 9–11 is the same as (A) of figure 9–11 except that the current to the load is not rectified, therefore the load receives a-c power only.

Both circuits make use of a feedback winding. These circuits are similar in that both provide a direct current feedback. This is accomplished through the use of such

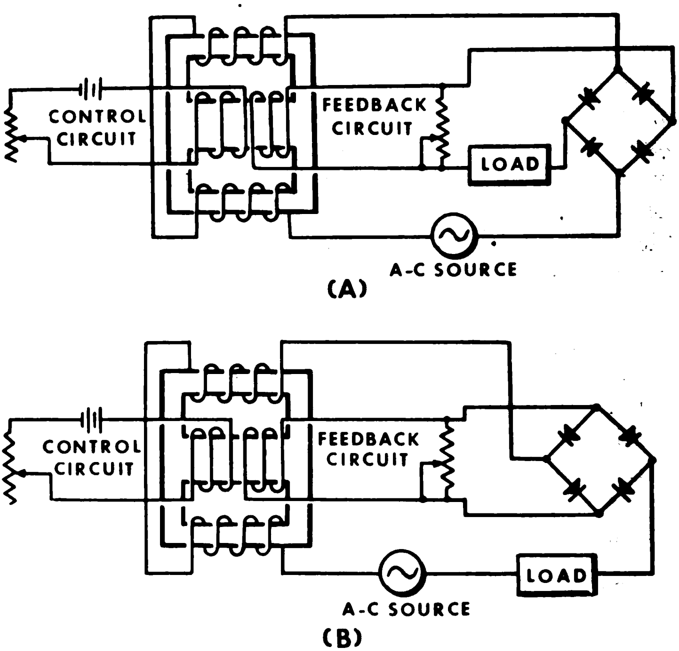


Figure 9–11.—A magnetic amplifier employing feedback.

circuit arrangement that the output always flows through the feedback winding in the same direction.

The feedback can be positive or negative depending on the connections to the feedback winding. If the flux produced by the feedback winding aids the control winding flux, the feedback is positive (regenerative) ; if it opposes the control winding flux, it is negative (degenerative). Positive feedback makes the circuit more sensitive to changes in control current, thus, extremely high gains are possible. However, this type feedback has two disadvantages for it increases the time-lag characteristics of the amplifier and may also cause instability. While negative feedback increases the linearity of the magnetization curve and decreases the amplifier's time-lag characteristics, it also reduces the sensitivity of the amplifier.

The curves in figure 9–12 illustrate the effect of both positive and negative feedback on load current as a function of input control current. It should be noted that the same amount of d-c control current is needed to saturate the core regardless of the amount of feedback provided. The reason that less input control current is necessary with positive feedback is that the feedback current adds to the control winding current. As illustrated in figure 9–12, the difference between the control current with positive feedback *A* and the control current without feedback *B* is actually supplied by the output circuit by

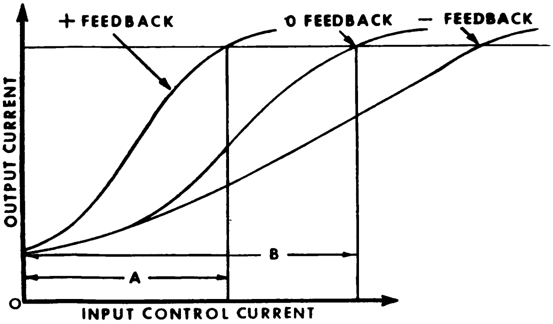


Figure 9–12.—Effects of external feedback.

means of the feedback winding. In both of the circuits illustrated in figure 9–11, the load current flows through the parallel combination of the feedback winding and a potentiometer. The positioning of the potentiometer provides a means of controlling the amount of feedback current.

Figure 9–12 also shows the effect of negative feedback in a magnetic amplifier. The negative feedback curve indicates that the input control current has less control of the output current, but that the response of output control current to input control current is very linear.

The TIME OF RESPONSE, or delay in operation, after a signal is applied is a serious disadvantage in some applications of magnetic amplifiers. The ordinary saturable reactor type, or saturable transformer, requires a certain time for a control action to take effect since the action is applied to an L–R control circuit. The time of response is usually proportional to the time constant of the d-c input control winding. The time constant for this inductive circuit may be determined by the equation

$$t = L/R$$

where

t is in seconds

L is inductance in henrys of the control winding

R is the resistance in ohms of the control winding.

This time constant may be reduced by increasing the control circuit resistance, but this causes a control power loss. Degenerative feedback can be employed to do the same, if the reduced gain can be tolerated. A general arrangement would be to apply degenerative feedback and add more stages in cascade. When stages are cascaded, a d-c output of the first stage is applied to the control winding of the next stage, and so on.

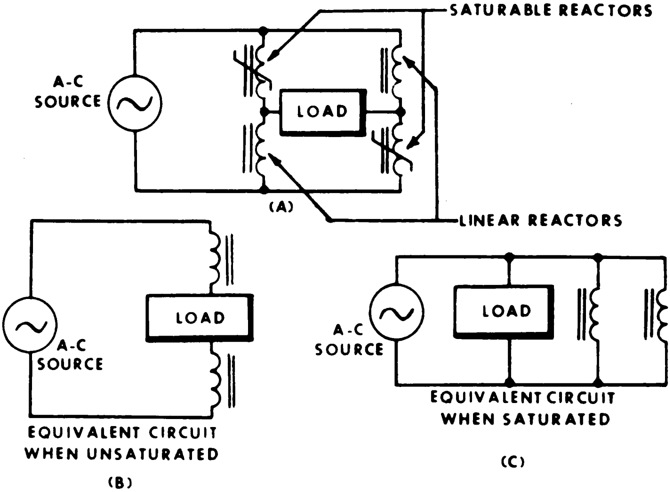
In magnetic amplifier circuits the speed of response is mainly determined by the time of one half cycle of the a-c supply frequency. This time can be made quite small by using a higher a-c supply frequency such as the 400-cycle supply in an aircraft.

One result of the time delay is that the frequency of the control signal must be lower than that of the a-c source. This is a severe limitation as compared with vacuum-tube amplifiers, in which the maximum frequency is limited only by interelectrode capacities and transit time.

Time delay increases with amplification and in high gain magnetic amplifiers becomes a limiting factor in their practical application. At fairly high gains the signal frequency is limited to about 10 percent of the a-c source frequency. At very low gains the signal frequency can be more than 50 percent of the a-c frequency. The source frequency can be as high as can be obtained with suitable choice of available core materials. Frequencies in megacycles have been used to energize magnetic amplifiers.

Magnetic amplifiers are used in many applications where it is necessary to use higher frequency power sources than are available from conventional supplies. For example, when magnetic amplifiers are used for audio amplification, it is necessary to generate a carrier frequency at least three times the highest audiofrequency. Vacuum-tube oscillators and rotating machines are unsuitable for this function because they introduce short life or moving parts, thus losing the major advantages of magnetic amplifiers.

An answer to this problem of higher frequencies has been the development of FREQUENCY MULTIPLYING DEVICES that operate on the saturable reactor principle. These devices utilize iron-core transformers in somewhat the same manner as magnetic amplifiers. Frequency multipliers of this type operate on the principle that unsymmetrical variations of magnetic flux are produced by an alternating current when it is acting upon a magnetic core which is already saturated by a d-c flux. In this arrangement,



EQUIVALENT CIRCUIT WHEN UNSATURATED

(B)

EQUIVALENT CIRCUIT WHEN SATURATED

(C)

Figure 9–13.—Bridge type frequency multiplier (tripler).

the currents that are induced in the secondary windings may be twice the frequency of the primary current as a result of the core's saturation.

It is also possible to construct devices that increase the frequency threefold. These are known as triplers and operate somewhat similar to the doublers. Frequency tripling is accomplished by combining a sinusoidal and a distorted voltage. A bridge type frequency tripler is shown in figure 9–13 (A).

In this circuit, linear reactors are used to produce the sinusoidal voltage and the necessary phase shift. The saturable reactors produce the distorted waveform and also the switching action that will be described later. If the impedance of the saturable reactors is zero when saturated and infinite when unsaturated, the equivalent circuits of the tripler may be drawn as shown in (B) and (C) of figure 9–13. This circuit utilizes two linear reactors and two saturable reactors. The saturable reactors become saturated at near peak amplitude of the source current while the linear reactors never become saturated.

When the saturable reactors are unsaturated (offer infinite impedance), the circuit is equivalent to that shown in (B), for the saturable reactors are effectively out of the circuit. When the reactors are saturated they offer zero impedance, thus the load is effectively connected across the a-c source potential. Figure 9–13 (C) shows an equivalent of the bridge circuit during saturation.

By referring to the circuit in (A) it can be seen that when the top terminal of the source voltage is positive the current will flow through the load from left to right if the saturable reactors are unsaturated. When the reactors are saturated their impedance is practically zero and the current will flow from right to left through the load. It may be noted that through the saturation of the reactors, which occurs during a portion of each half cycle of the a-c source, the direction of load current will be reversed without a reversal of the a-c source potential. This switching action effectively changes the circuit alternately between the two equivalents shown in (B) and (C). In the circuit in (C), the load voltage is in phase with the line voltage. The linear reactors in the circuit in (B) cause the load voltage to lag the line voltage. The switching action which occurs during each half cycle produces an alternating current through the load, whose frequency is three times that of the a-c source. This threefold increase in frequency is a result of the circuit action which alternately switches the load across two out-of-phase voltages.

A further increase in frequency may be obtained by connecting multiplier units in cascade. For example, three doublers connected in cascade will produce a frequency eight times the original frequency. It is also possible to connect doublers and triplers in the same circuit for purposes of frequency multiplication.

Applications of Magnetic Amplifiers

The magnetic amplifier has found application in almost all circuits which use vacuum tubes and many which do not. Among such circuits are amplifiers, regulators, servo systems, counting and dividing circuits, and many high frequency uses. The application of the magnetic amplifier to several of these categories will now be described.

AUDIOFREQUENCY AMPLIFIER.—Figure 9–14 (A) shows the schematic of an audiofrequency magnetic amplifier. Part (B) shows a simplified schematic of the circuit in (A). It is given in order that you may become familiar with this type of presentation which is frequently used in many publications dealing with magnetic amplifiers.

As previously explained, in connection with frequency multipliers, the a-c source potential must have a frequency somewhat greater than the highest audiofrequency to be amplified. In this circuit the audio signal is applied to the control winding of the magnetic amplifier and thus replaces the d-c control source. (See fig. 9–11.)

Magnetic amplifiers used in audiofrequency applications are usually operated in class AB push-pull because class A operation would not only be inefficient but also provisions would have to be made to separate the carrier from the voice frequency.

In figure 9–14, two control windings are used—a bias winding and a modulator winding. The d-c current through the bias winding provides the required directional flux to set up the desired point of operation on the magnetization curve. With no audio modulation applied, there are equal and opposite load voltages on the primary of the output transformer, resulting in zero output.

When audio modulation is applied, it will act as a source of external feedback, reinforcing the directional flux in one leg of the coil and opposing it in the other leg. This will cause the a-c current flowing through the two legs to vary at the impressed audiofrequency, this variation being opposite in each leg; that is, as the current through one leg is increasing, the current through the other is decreasing. This will cause an increased audio voltage across one-half of the output transformer

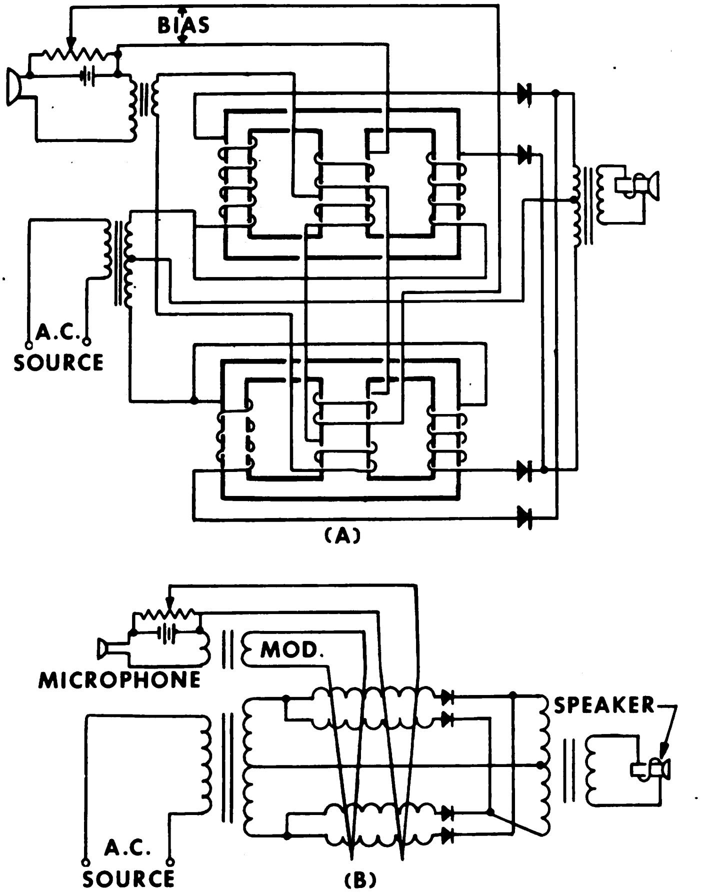


Figure 9–14.—Audiofrequency magnetic amplifier.

primary and a correspondingly decreased audio voltage drop across the other. Since the voltages on the output transformer primary are no longer equal and opposite, a resultant voltage, which is an amplified form of the modulating voltage, appears on the secondary of the output transformer, thereby actuating the speaker. The voltages and currents present at the various parts of this circuit are shown in figure 9–15.

SERVO SYSTEMS.—The most common uses of magnetic

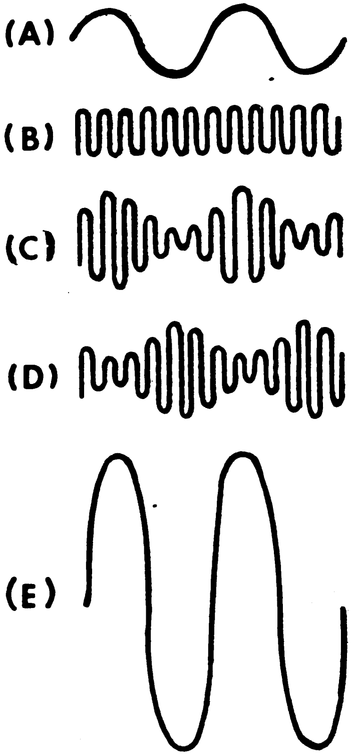


Figure 9–15.—Voltages and currents present at various parts of the audio amplifier circuit: (A) Modulating voltage; (B) supply voltage input to a-c line; (C) current in upper leg or voltage in upper half of output transformer primary; (D) current in lower leg or voltage in lower half of output transformer primary; (E) output voltage to speaker.

amplifiers in aviation today is in connection with servo systems. In order to illustrate some applications of the principles of magnetic amplifiers that have been presented in this chapter, three circuits will be described. The first two circuits utilize magnetic amplifiers only in the output stages. The third circuit is a complete servo system that uses only magnetic amplifiers.

Figure 9–16 illustrates an interesting type of servo amplifier in that magnetic amplifiers provide the voltage for both phases of a 2-phase induction motor. It is the normal practice in servo amplifiers to connect one of the

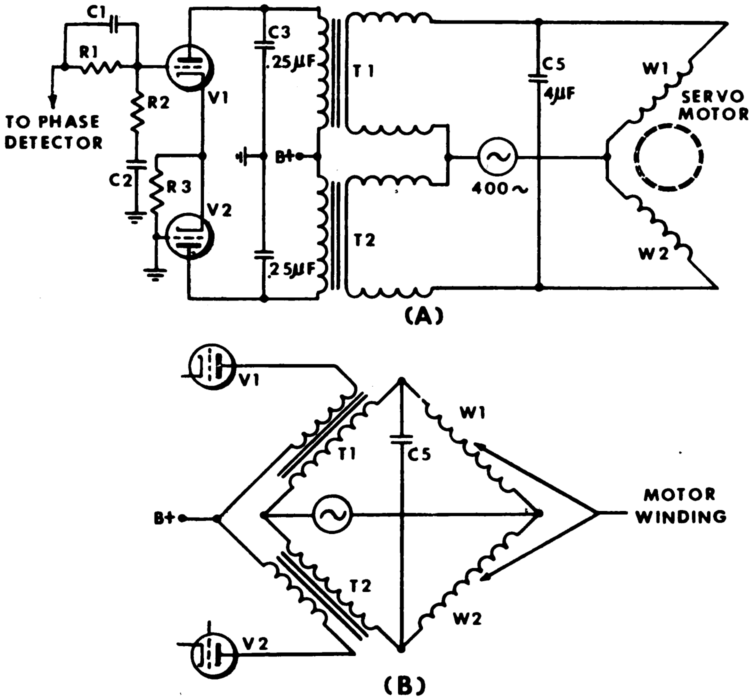


Figure 9–16.—Magnetic amplifiers used to control a 2-phase induction motor.

motor windings (fixed voltage winding) across the a-c source in series with a capacitor in order to provide the required 90-degree phase shift; the other winding (controlled winding) is connected to the amplifier's output.

The input to the amplifier of figure 9–16 (A) is the d-c output of a phase detector. This input is applied to the grid of *V*1 through a d-c lead network consisting of *C*1 and *R*1. The R–C combination consisting of *R*2 and *C*2 which is connected between the grid of *V*1 and ground is a d-c retard network. These networks simply serve as antihunt devices of a different type than the mechanical friction and the electrical damping antihunt systems.

Tubes *V*1 and *V*2 constitute a cathode-coupled paraphase amplifier working into two saturable reactors (magnetic amplifiers). With zero signal, both tubes draw an equal amount of plate current and the voltages applied to the motor windings are in phase, thus there is no motor rotation. The saturable reactor secondaries and the motor windings form an a-c bridge circuit. The equivalent circuit shown in figure 9–16 (B) is a redrawing of the circuit in (A) and illustrates the bridge circuitry. The secondaries have a large impedance when the control current is small and practically zero impedance with a large control current. (Control current varies the permeability of the core.)

With reference to figure 9–16 (B), it may be seen that when *V*1 is conducting heavily *V*2 will be near cutoff. At this time the secondary inductance of *T*1 will be practically zero thus motor winding *W*1 will be effectively connected across the a-c source. Also, at this time the secondary inductance of *T*2 will be very large (effectively an open circuit), thus, motor winding *W*2 is effectively connected across the a-c source in series with capacitor *C*5.

In a similar manner, when *V*2 conducts heavily, *V*1 will be near cutoff, the secondary inductance of *T*2 will be near zero, and the secondary inductance of *T*1 will approach maximum. Under these conditions motor winding *W*2 is now effectively connected across the a-c source and *W*1 is connected across the source in series with capacitor *C*5. The motor will now rotate in the opposite direction.

From the explanation just given it can be seen that the speed and direction of the motor can be controlled by the amplitude and phase of the input signal.

The second circuit to be described illustrates the use of feedback windings and rectifiers in order to obtain maximum sensitivity. In figure 9–17 the magnetic amplifier *T*1 is the power output stage of the servo amplifier. By means of the d-c output current from *V*1 and *V*2 it controls the magnitude and phase of the 400 c.p.s. a-c voltage fed to the controlled winding of the servomotor,

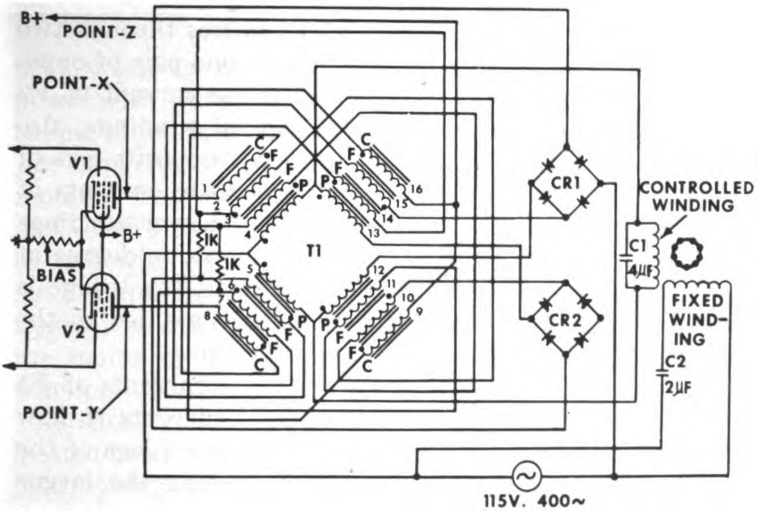


Figure 9–17.—Magnetic amplifier using feedback.

thus controlling its speed and direction. The magnetic amplifier accomplished this control basically by varying the reactances, and hence the a-c voltage drops, of the power windings *P* connected in series with the 400 c.p.s. supply voltage and the controlled winding of the motor. The variation of these inductances is achieved by changing the degree of magnetic saturation of the core by passing direct current through a control winding *C* on the same core with each of the power windings. This direct current is controlled by the d-c error voltage of the servo amplifier. Additional saturation current obtained by external feedback is supplied to extra control windings *F* in *T*1 in order to obtain more positive control.

The power windings of *T*1 are arranged in a bridge circuit which includes bridge rectifiers *CR*1 in one leg and *CR*2 in the adjacent leg. Associated with each power winding is a control winding on the same core. The 400 c.p.s. a-c supply is connected across one pair of opposite corners (left and right) of the bridge (input) and the controlled field winding of the motor receives voltage from the other pair of opposite corners (output) of the bridge. The d-c plate current of V1 passes through two control windings connected in series in one pair of opposite legs of the bridge, and the d-c plate current of V2 passes through the other pair of control windings, also connected in series, in the other pair of opposite legs of the bridge. With no error signal, the plate currents of V1 and V2 are equal, and therefore the power windings all offer the same reactance because the same degree of saturation exists in each core. Therefore, the bridge is balanced and the bridge output voltage applied to the controlled field winding of the motor is zero.

A d-c error signal will make the plate currents of V1 and V2 unequal, thus causing unequal currents to flow through the control windings in the adjacent legs of the bridge. The cores of the windings passing the larger current will saturate to a greater degree than before, thus reducing the reactance of the associated power windings. The other two cores will saturate less and therefore the reactances of these power windings will increase. Thus the bridge will become unbalanced and the controlled field winding of the motor will receive a voltage from the output of the bridge. The motor will then rotate in such a direction as to reduce the error and again balance the bridge.

The magnetic amplifier utilizes a core material exhibiting a large change in magnetization with a relatively small change in magnetizing force. As a result, a comparatively small d-c control current will exert positive control of the motor. Additional sensitivity is secured by passing current obtained from the output of bridge rectifiers CR1 and CR2 through extra windings on each core. With this positive external feedback arrangement a current is supplied through the feedback windings which is in phase with that flowing through the corresponding control winding. This produces a given saturation with less control current, thus providing more effective control.

The 2-phase motor has a fixed field and a controlled field which operate from a-c voltages 90 degrees out of phase with each other. Reference voltage from the a-c supply is fed to the fixed winding through phase modifying capacitor *C*2. This capacitor in series with the motor's fixed voltage winding establishes the proper phase and increases the amplitude of the voltage applied to this winding. The output waveform from the magnetic amplifier is improved and the proper phase relationship is maintained between the two field voltages of the motor by *C*1 in shunt with the controlled voltage winding.

It should be noted that the dots (fig. 9–17) that are located near one end of each core indicate polarity or phase relationship between the windings on each core. When the dots are located at the same end those windings are in phase; and if the dots are at opposite ends, the windings are out of phase.

The outputs from the bridge rectifiers are from the top and bottom points. By tracing the circuit from one of the output points through the feedback windings to the opposite output point it may be seen that the feedback current will flow through a winding on each of the four cores. When tracing this circuit it may be noted that the polarities of the feedback windings are such that they are in phase with the control windings that are decreasing the permeability of their core. The other feedback windings are out of phase with the control windings that are increasing the permeability of their cores; thus, by being out of phase these feedback windings actually aid their control windings which are tending to increase the permeability.

The third circuit to be described is a complete servo amplifier that does not make use of vacuum tubes. The circuit is illustrated in figure 9–18 and consists of a demodulator, voltage amplifier, and power amplifier. In order to use magnetic amplifiers with synchros in a servo system, it is necessary to convert the alternating current error signal into a direct current control signal. This

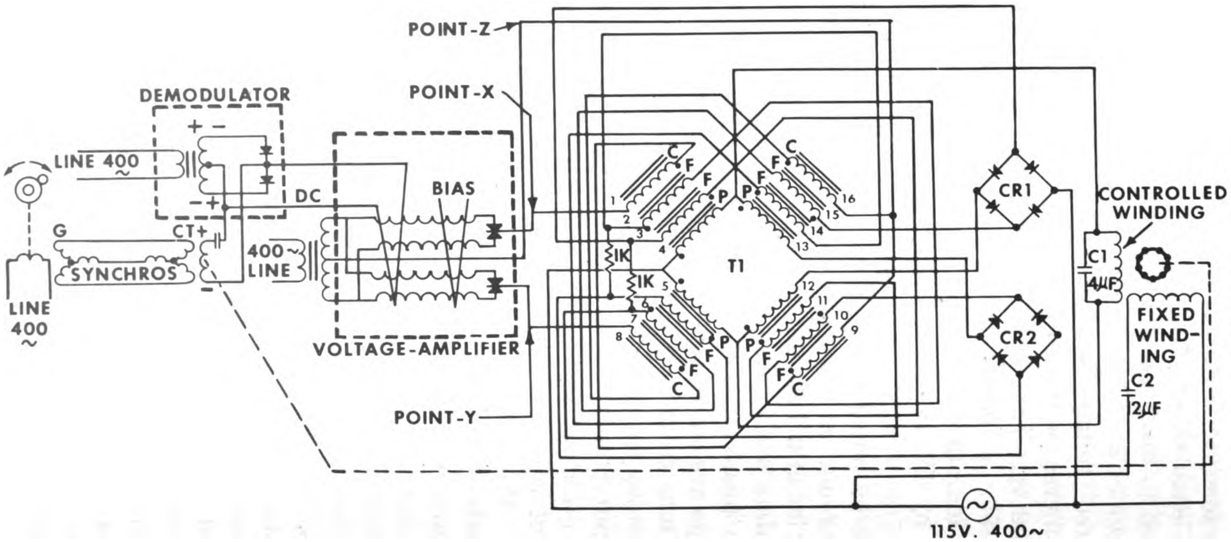


Figure 9–18.—Complete magnetic amplifier servo.

function is performed by the demodulator (phase sensitive rectifier) and its operation is somewhat similar to that described in chapter 8 in connection with figure 8–5, except that dry rectifiers are used in place of vacuum-tube diodes.

The output of the demodulator is applied to the control windings of the voltage amplifier. This stage consists of four reactors, each of which has a control winding and a bias winding. The voltage amplifier stage is similar in operation to the circuit shown in figure 9–10 (A), for it is actually a combination of two magnetic amplifiers each of which is similar to the one in the referenced figure. The only exceptions are that a bias winding has been added and the output from each amplifier is d.c. due to the different rectifier connections.

The two outputs from the voltage amplifier are coupled to the power amplifier stage which is identical to the amplifier shown in figure 9–17. Since this amplifier has been described in detail, such will not be repeated. However, it should be noted that in figure 9–17 tubes *V*1 and *V*2 regulated the current through the control windings, while in this circuit the control current is now the outputs of a magnetic amplifier. Points *X*, *Y*, and *Z* have been identified in both of these figures in order to aid in showing how a magnetic amplifier may be used to replace vacuum tubes as a driver for the power amplifier.

ELECTRONIC POWER SUPPLY REGULATOR.—Many aviation electronic equipments contain voltage regulated power supplies. The accuracy of the voltage produced by these supplies is important in so far as proper operation and maximum life of the equipment is concerned. Present day electronic equipments contain many timing, waveshaping, and other critical circuits that require a constant d-c voltage source.

Figure 9–19 (A) shows a power supply which uses a magnetic amplifier to regulate the d-c output voltage. In this circuit the conventional filter choke has been replaced by a saturable reactor *T*2. The voltage regulator (VR) tube *V*1 produces a constant reference voltage across coil *B* and acts as a current amplifier for the control coil winding *A*. Coils *A* and *B* control the saturation of *T*2, hence, the inductance of the load winding. This action effectively transfers filter capacitor *C* to the other side of the choke when the core is saturated. The end result is that the power supply alternates between a choke input and a capacitor input type of supply, thus a constant voltage output is produced.

Figure 9–19 (B) shows a variation of figure 9–19 (A) in which saturable reactor *T*2 is connected in series with the primary of power transformer *T*1. An advantage of this circuit arrangement is that a regulated filament voltage is also produced.

The significance of constant filament voltage is appreciated when it is realized that an estimated 80 percent of all electronic failures can be directly traced to tube or capacitor failures—this does not include transformer, choke, and resistance failures which can usually be traced to overloads occurring during these failures.

Tubes are listed as the greatest single cause of electronics failures. Cathodes or filaments have an optimum temperature of operation. The incident of failure rises rapidly as the temperature varies around this optimum. Very little data exist on the effect of under-temperature operation. It is known that although the latter condition is not as serious as the other, it does shorten tube life. It is desirable to operate the heaters at normal rating from regulated power supplies when long stable service is required of tubes. Life tests on tubes operated on above normal voltages are not available; however, life tests on incandescent electric lights show that the life of the filament is reduced 50 percent when operating on a voltage only 7 percent above normal.

In electron tubes a rise in line voltage also causes an increase in plate current which adds to the abuse of the already overheated filament or cathode. The additional heat generated by the tube and other components within

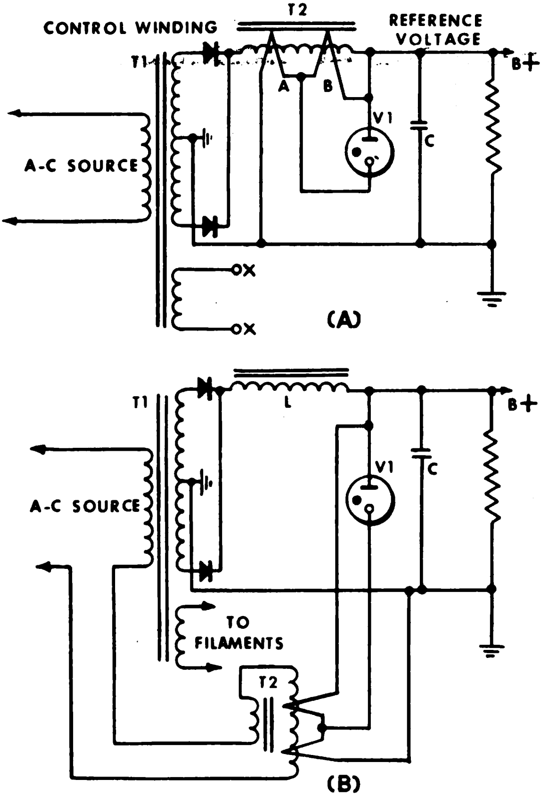


Figure 9–19.—Power supply using magnetic amplifier voltage regulator.

the equipment compartments cause the ambient temperature to rise which also contributes somewhat towards accelerated equipment failures.

Another point often overlooked is that once the tubes have been operated under accelerated conditions their performance is considerably reduced when returned to normal operating voltages. Military equipment is designed to operate at about ±5 percent of a specified voltage. The service free life however is reduced considerably when operating in the upper voltage range.

VOLTAGE AND FREQUENCY REGULATORS FOR ALTERNATORS AND INVERTERS.—Magnetic amplifiers are being used to provide a rugged, reliable, long-lived device to maintain constant voltage output from aircraft alternators. Regulators which use magnetic amplifiers have no moving parts (except for relays) thus they are unaffected by acceleration or severe vibration. They are long-lived, with expected trouble-free life in excess of 5,000 hours.

Figure 9–20 shows a functional diagram of a voltage regulator of this type. The constant voltage reference is provided by a VR tube. Although this regulator is equipped with two VR tubes which are connected in parallel, only one functions at a time. The reason that only one tube functions is that no two VR tubes have the same ionization potential. Since this is true, the tube with the lower firing potential will ionize first and hold the voltage below the ionization potential of the other tube. However, should the ionized tube become inoperative the second tube will ionize and provide the constant voltage reference.

The first magnetic amplifier stage consists of two small saturable reactors, connected push-pull. Each has its control winding connected in series opposition to the voltage comparator bridge. Each winding controls the current flowing from the a-c generator through the reactor's output winding.

When a-c generator output voltage is at the selected value, there is no error signal, and balanced output will flow from the two first-stage amplifiers. This first-stage output is rectified and applied to the control windings of the second-stage saturable reactor. The output winding of one of the first stage amplifiers is connected to the "buck" control winding in the second stage; the other is connected to the "boost" control winding. If a-c generator voltage is correct, equal currents will flow from the first stage through the "boost" and "buck" windings of the second stage.

If a-c generator voltage varies from the selected value, the error signal applied in the first stage will cause it to increase the flow of current in the appropriate ("buck"

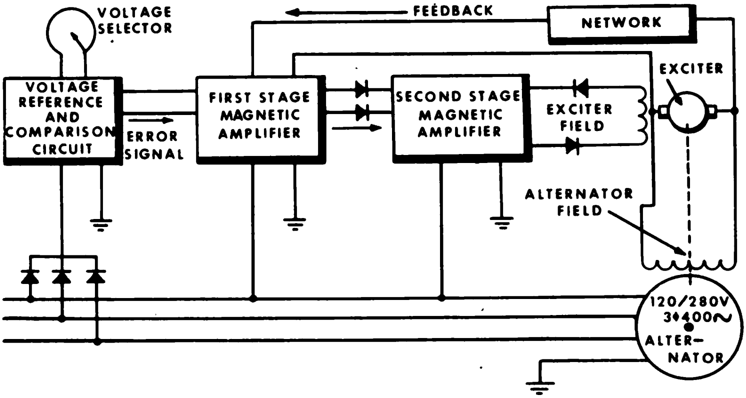


Figure 9–20.—Alternator type magnetic amplifier voltage regulator.

or "boost") second-stage control winding. This in turn increases or decreases the output current from the second stage, which is then rectified and applied to the main field of the exciter.

The exciter is a small d-c generator which is built into the a-c generator and driven from the same shaft. It provides the relatively high-power level of excitation for the main field of the a-c generator. In a sense, the exciter is another stage of amplification to raise the power level of the output from the voltage regulator.

The polarity of the error signal from the bridge comparator is determined by whether the a-c generator voltage has risen above or fallen below the selected value. This signal determines whether the push-pull stage passes the greater current through the "boost" or the "buck" windings of the second stage. This in turn determines whether the exciter field current is increased or decreased, thus controlling a-c generator field excitation and hence its output voltage.

The voltage regulator is a "closed-loop" system in the sense that a regulator change in exciter field current causes a change in alternator voltage which will in turn cancel the initial error signal calling for the change in field current. However, this closed loop includes a feedback circuit to increase system stability and limit voltage overshoot. The feedback circuit permits the regulator to anticipate the pending change in a-c generator output voltage.

The a-c generator exciter has an extra field winding (not shown in figure 9–20) which functions as a biasing device. It is excited from rectified a-c generator output and acts to partially cancel the main exciter field. This allows the second-stage reactor to supply sufficiently high currents to operate on the proper portion of its saturation curve. Because of the special exciter bias field required, the static regulator cannot be easily substituted for existing carbon-pile regulators in a-c systems.

Figure 9–21 illustrates the principle of applying the magnetic amplifier to FREQUENCY AND VOLTAGE CONTROLS of an aircraft inverter. A saturable reactor is in series with the field of the drive motor. This reactor controls the field current of the drive motor and thus its speed. The saturable reactor has two control windings; the current through these windings is controlled by two tuned circuits that are resonant at 350 and 450 cycles. The lower frequency coil is wound aiding to increase the field current, hence it may be seen that as the frequency of the output voltage drops the motor's field current will increase the speed of the inverter and thus the frequency of the output voltage. The higher frequency coil is wound opposing and causes the opposite effect. The output voltage is controlled by a second saturable reactor which regulates the generator's field current.

It may be noted from the response curves at the bottom part of the figure that if the output frequency varies above or below 400 cycles, increased current flows through the control winding connected to the resonant circuit whose frequency is nearer the frequency of the output voltage. The reaction of the two currents maintains a constant 400-cycle output.

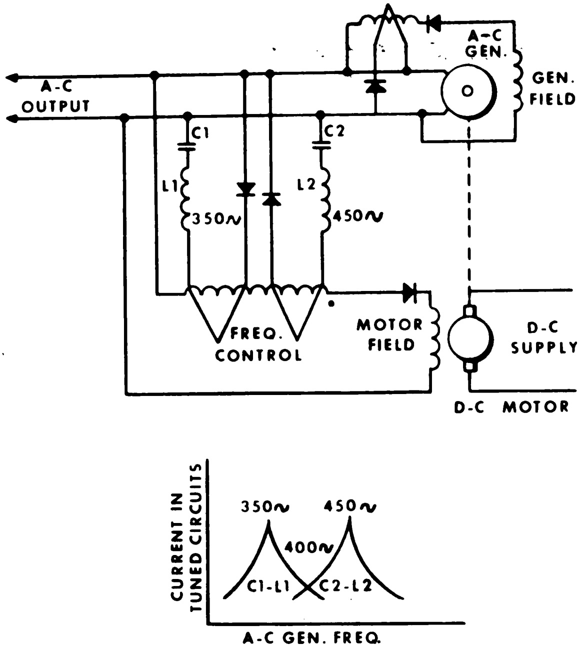


Figure 9–21.—Inverter—voltage and frequency regulated by magnetic amplifier.

Other applications of magnetic amplifiers:

1. TIMERS—Timing pulse generator for radar and loran equipment.
2. ELECTRICAL COMPUTERS—Add, subtract, multiply, divide, differentiate, or integrate electrical quantities.
3. MODULATORS—Voice frequency modulators and RF carrier trigger circuits.
4. MAGNETOMETER AMPLIFIERS—For magnetic air detection systems.
5. MULTIVIBRATORS—One shot and free running square wave generators.
6. SWEEP GENERATOR—Trigger and pulse forming circuits for radar and loran equipment.

7. COUNTERS, DIVIDERS, AND WAVESHAPERS—For loran and radar indicators.

ADVANTAGES AND DISADVANTAGES OF MAGNETIC AMPLIFIERS.—The advantages and disadvantages of magnetic amplifiers have been covered somewhat in detail in appendix I of *AT 3 & 2,* NavPers 10317. Since this information is available, it will not be repeated in this training course.

A summary statement as to the advantages of magnetic amplifiers is: They have long life, are rugged, can be hermetically sealed, no warmup time is required, there is low heat dissipation and low temperature rise, no high voltage d-c supply is required, the circuits are quite simple, are capable of high gain and can control large output power, and are characterized by ease of maintenance. Their major disadvantages center around the facts that there is a long timelag due to the inductive circuits, higher frequency supplies are necessary, and the output waveforms are not true sinusoids.

QUIZ

1. Basically, what is a magnetic amplifier?
    a. A variable load in series with an a-c source
    b. A load and an a-c source in parallel with a variable inductance
    c. A controlled variable inductance in series with an a-c power source and a load
    d. A variable inductance in parallel with an a-c power source and a load

2. How is it possible to control the inductance of a given coil?
    a. By controlling the permeability of the core
    b. By increasing the current flow in the coil
    c. By increasing the voltage applied to the coil
    d. By controlling the frequency applied to the coil

To answer questions 3, 4, and 5, refer to figure 9–1.

3. When will the inductive reactance of the coil be maximum?
    a. When the iron core is fully inserted into the coil
    b. When the a-c source voltage is maximum positive
    c. When the iron core is fully retracted from the coil
    d. When the a-c source voltage is maximum negative

4. When the iron core is fully retracted from the coil, the voltage drop across the output load is
    a. minimum a.c.
    b. zero
    c. maximum d.c.
    d. maximum a.c.

5. What is the disadvantage of using this simple magnetic amplifier?
    a. Power requirements are excessive.
    b. It is a mechanical device.
    c. It has no low frequency limit.
    d. It has no top frequency limit.

6. An electronic method of controlling the permeability of a coil core is obtained by
   a. decreasing the number of turns on the coil
   b. increasing the number of turns on the coil
   c. using a properly designed magnetic core material
   d. saturating the core with a d-c or properly phased a-c voltage

7. A saturated core will offer how much impedance to the a-c source applied?
   a. Maximum
   b. None
   c. Minimum
   d. Zero

8. On the magnetization curve shown in figure 9–2 where is the point of operation that will give the desired magnetic amplifier action?
   a. Point *b*
   b. Point *a*
   c. Point *d*
   d. Point *c*

9. What is the advantage of using a three-legged core saturable reactor in preference to using a two-legged one?
   a. It will eliminate transformer action between the secondary winding and the control winding.
   b. It is adaptable only to saturable reactors.
   c. There will be more current drawn from the a-c source.
   d. A-c voltage will be induced into the control winding.

To answer questions 10 through 13, refer to figure 9–6.

10. Where must the control pot be positioned to give maximum saturation of the core material?
   a. At the negative end of the voltage divider
   b. At the midpoint on the voltage divider
   c. Between the midpoint and the positive end of the voltage divider
   d. At the positive end of the voltage divider

11. As shown, the output across RL is
   a. zero
   b. intermediate
   c. maximum
   d. minimum

12. When will there be minimum voltage across RL?
    a. When the control pot is at the top of the voltage divider
    b. When the control pot is at midposition
    c. When the control pot is at the bottom of the voltage divider
    d. When the control pot is between midpoint and the bottom of the voltage divider

13. The circuit shown in figure 9–6 has two distinct periods of operation, and they are known as the
    a. control period and the negative period
    b. reset period and the control period
    c. operating period and the reset period
    d. operating period and the control period

14. What is the effect of using external positive bias in magnetic amplifiers?
    a. Decreased gain and decreased time of response
    b. Increased gain and decreased time of response
    c. Increased gain and increased time of response
    d. Decreased gain and increased time of response

15. When TIME OF RESPONSE is a design consideration, what type feedback bias could be employed?
    a. External negative bias
    b. External positive bias
    c. Regenerative bias
    d. Inphase feedback

16. The action of rectifiers in series with the load current produces what type of feedback?
    a. Negative feedback
    b. External feedback
    c. Internal feedback
    d. Degenerative feedback

17. In self-saturating type magnetic amplifiers, what determines the speed of response?
    a. The time of one-half cycle of the a-c supply frequency
    b. The resistance of the load
    c. The inductive reactance of the control winding
    d. The turns ratio of the control winding

18. The time of response of a magnetic amplifier may be determined by which of the following?
    a. $t = R/L$
    b. $t = RL$
    c. $t = 2/LR$
    d. $t = L/R$

19. Vacuum-tube oscillators are unsuitable for generating a carrier frequency for an audio magnetic amplifier because the
    a. short life of vacuum tubes causes loss of the major advantages of magnetic amplifiers
    b. transit time of vacuum tubes is a limiting factor for upper frequencies
    c. upper frequency limit of magnetic amplifiers is too low for vacuum-tube oscillators
    d. vacuum-tube oscillators may drift thus moving the operating point on the hysteresis loop to a more unstable point

20. Frequency multipliers operate on the principle of
    a. negative feedback caused by the proper phasing of the audiofrequencies
    b. unsymmetrical variations of magnetic flux produced by an alternating current acting upon a saturated core
    c. symmetrical variations of magnetic flux in an unsaturated magnetic core caused by an alternating current
    d. degenerative feedback decreasing the time of response thus increasing the frequency

21. (Refer to figure 9–16.) *V*1 goes into maximum conduction causing
    a. *T*1 to saturate, thus placing *C*5 in series with ***W*1**, causing the motor to turn
    b. *T*2 to saturate, thus placing *C*5 in series with ***W*2**, causing the motor to turn
    c. *T*1 to saturate, thus placing *C*5 in series with ***W*2**, causing the motor to turn
    d. *T*2 to saturate, thus placing *C*5 in series with ***W*1**, causing the motor to turn

CHAPTER 10

INTRODUCTION TO TRANSISTORS

The transistor is a relatively new device that uses the unusual properties of semiconductors to perform many of the functions formerly possible only with vacuum tubes. Transistors may be used in circuits to provide amplification, oscillations, pulse generation and counting, timing and gating, and many other uses. The term transistor is coined from "TRANSfer" and "resISTOR."

The transistor is much smaller, lighter, longer lived, more rugged, fantastically more efficient, and potentially less costly than the vacuum tube. Furthermore, little current is used in its operation and practically no heat is generated, thus little space is required. These qualities will probably make the transistor extremely useful in the construction of portable and airborne electronic equipment. It offers the additional advantage of instant warm-up, thus no standby power is required.

The technician may expect a gradual introduction of the transistor in equipment he is called on to repair and maintain. Some equipments already utilizing transistors are hearing aids, portable record players, radios, electronic megaphones, aircraft voltage regulators, aircraft intercommunication systems, and portable test and measuring equipment. The military uses of the transistor are obvious, although few details have been officially released. It is potentially suited for use in guided missiles, proximity fuzes, radio and radar equipment, remote controlled devices, miniature teletype and communication equipment, and compact airborne computers. In short, they are used wherever light weight, small size, and low power consumption are desired.

The major objective of this chapter is to introduce the basic principles of transistor operation. New concepts are needed in order to understand its somewhat unexpected performance. The theories associated with transistors are radically different from those associated with vacuum tubes. Unless you feel that you are familiar with the basic theories of atomic structure it is suggested that you review chapter 1 of *Basic Electricity,* NavPers 10086.

SEMICONDUCTORS

Transistors represent only a part of a large group of semiconductor devices. Included with transistors in this large grouping are the germanium diodes which are used as detectors, discriminators, clippers, and limiters; and the silicon diodes used as high-frequency mixers.

The old "crystal" radio receiver, employing a galena crystal and an adjustable "cat whisker," represents one of the earliest applications of semiconductor devices to radio.

The operation of transistors, as well as the diodes, depends upon the electrical properties of a class of substances known as semiconductors. A semiconductor is a solid material which possesses a conductivity greater than a good insulator but considerable less than that of a good conductor. Under some conditions it acts like an insulator (such as glass).

Silicon and germanium, when combined with certain impurities, are the most popular semiconductors. There are other materials exhibiting similar properties but they are not used as extensively as the two mentioned. The principles which determine the behavior of germanium and silicon can easily be applied to these materials. Because of these facts, this chapter will deal only with silicon and germanium.

ELECTRONS AND HOLES IN SEMICONDUCTORS.—The new concept associated with transistors is that there are two characteristically different methods of current flow through a semiconductor. One current is the familiar electron current, which is a drift of free or "excess" electrons under the influence of an electric field. This current is called EXCESS conduction, conduction by excess electrons, or simply conduction by electrons.

The second current, called DEFICIT conduction or conduction by holes, occurs in materials in which the valence bond between atoms is weakened by a deficiency of electrons, leaving a "hole" in the valence-bond structure. When an electric field (potential) is applied across such a material, the electrons jump from hole to hole, completing the valence-bond structure but resulting in the apparent motion of holes through the conductor in the opposite direction. An analogy to the apparent motion of holes is the motion of vapor bubbles up the liquid-filled columns on juke boxes. In these tubes a drop of liquid is vaporized at the bottom. As the liquid above the bubble moves downward to fill in the space produced, a bubble of vapor rises in the column. The rising bubble really is caused by downward motion of the liquid. Similarly, the motion of positively-charged holes is really motion in the opposite direction of negatively-charged electrons.

Because a hole is an absence of an electron, it is a region of positive charge. Because it moves, it is a current. Thus, you can think of "hole current" as being the motion of positive holes through the semiconductor. A brief review of the atomic structure of crystalline matter helps to explain conduction by holes. Crystalline matter is any substance in which the atoms are arranged in an orderly structure.

ATOMIC STRUCTURE.—An atom consists of a positive nucleus containing a fixed number of protons surrounded by an equal number of electrons in various orbits. Each orbit can contain not more than a fixed maximum number of electrons. When an orbit contains this fixed maximum number of electrons, it is said to be complete. In many atoms the inner orbits are complete and only the outer orbit is incomplete. The innermost (complete) orbits, together with the core, form a stable "ionic core" with a net positive charge. This core can be considered completely inactive as far as chemical reactions and electrical phenomena are concerned. Only the outermost ring of electrons determines the chemical and electrical characteristics of the substance.

CRYSTAL STRUCTURE.—Some of the outermost electrons form VALENCE BONDS between adjacent atoms. These bonds, which hold the atoms rigidly in place in crystalline substances, are stable for certain given numbers of electrons. The germanium atom has 32 electrons rotating around its nucleus. Twenty-eight of these electrons are tightly bound to the nucleus, while the remaining four are the electrons that enter into chemical reactions and are of importance to transistor physics. It should be noted that silicon is similar to germanium since they have the common property of being tetravalent—that is, only four of their outer electrons are able to enter into chemical reactions. Each of the four valence electrons of an atom form bonds with an electron from one of the four nearest atoms. These pairs of bonded electrons are commonly known as covalent bonds and provide the force that binds the atoms together to form a crystal structure—the crystalline form of silicon and germanium. The three-dimensional crystal lattice in figure 10–1 shows the covalent bonds as rods between the atoms. This figure is redrawn in two-dimensional form in figure 10–2, and shows how each bond between adjacent atoms is made of two electrons, one from each atom, resulting in a stable bond.

For simplicity of explanation, in connection with the two elements just mentioned, since the electrons in the inner shells are tightly bound to the nucleus, assume that the nucleus and tightly bound electrons are a core. This core has a positive charge of four, around which the remaining four electrons are rotating with a net negative charge of four. This condition is shown in figure 10–2 (A).

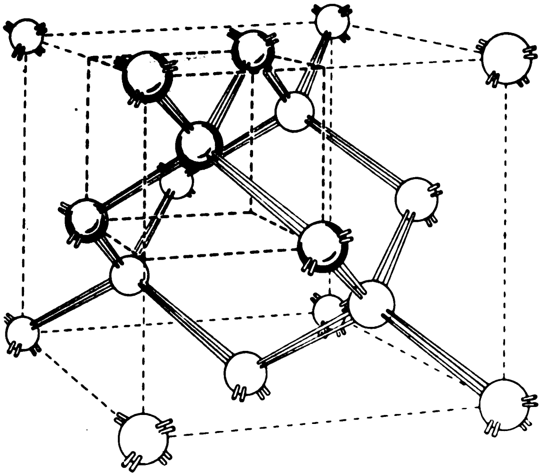
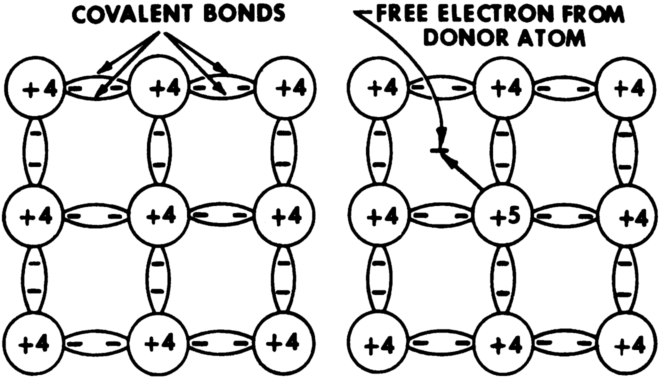


Figure 10–1.—Crystal structure showing covalent bonds.

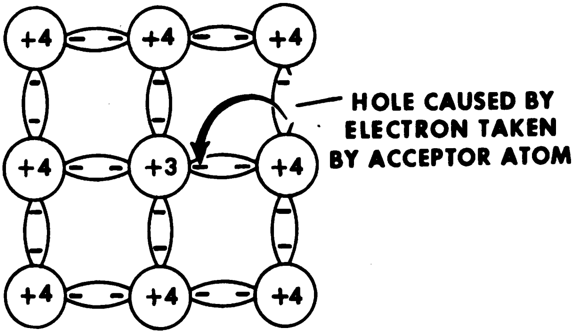
GERMANIUM CHARACTERISTICS.—The representation shown in figure 10–2 (A) is for pure germanium. Note that all atoms are equidistant from each other and that between the cores of any two atoms are two electrons. In the crystal structure of pure germanium a state of equilibrium exists, under which the force of attraction and repulsion between atoms and electrons exactly balance each other. Germanium in its purest state behaves like an insulator. This is true since there are no mobile carriers of electric current. However, due to thermal agitation, at ordinary room temperature, some electrons acquire sufficient energy to break their bonds and become free. These free electrons are able to move freely throughout the crystal structure and are available for conduction. As a result of this action, germanium (or silicon) is classified as a semiconductor rather than an insulator.

Almost all semiconductor devices permit electrical current flow to take place easier in one direction than in the other and thus may be used as rectifiers or diodes. In addition, other physical conditions may affect the flow of current. Such conditions include light, heat, and the



PURE GERMANIUM
(A)

GERMANIUM WITH DONOR
(B)



GERMANIUM WITH ACCEPTOR
(C)

Figure 10–2.—Germanium with and without impurities.

presence of other electrical fields. It is the last property that makes the transistor possible.

If a sample of germanium or silicon with broken bonds is placed in an electric field, two kinds of current can flow. The ejected electrons drift in the direction of the applied field, thus constituting an ordinary conduction current. The second current, however, results from the motion of electrons from one bond breaking away to fill up the hole caused by the absence of an electron from an adjacent bond. In other words an atom having a hole may take on an electron from an adjacent electrically neutral atom, thus leaving the second atom with a hole. The hole may thus migrate from one atom to another through the material. The hole effectively acts as if it were an electron with a positive charge; thus the action of the hole acts like a current flow of positively charged particles. These two currents are called ELECTRON CURRENT and HOLE CURRENT.

DONORS AND ACCEPTORS.—If an impurity such as arsenic is mixed with a semiconductor such as germanium or silicon, an application of a potential to the resultant mixture will produce more electron current than hole current. This results from the fact that each atom of arsenic with five valence electrons, when introduced into germanium or silicon as a low-fraction impurity, displaces one atom of germanium or silicon and forms four covalent bonds, using up four of its valence electrons. (See fig. 10–2 (B).) The extra electron has no adjacent electron with which to form a covalent bond, thus it is free to move within the crystal. A semiconductor with excess electrons, such as germanium with arsenic, is called an *n*-type substance because the current carriers are negative electrons. Arsenic, phosphorus, antimony, and other impurities, which make *n*-type semiconductors, are called DONORS.

If an impurity such as boron is added to the semiconductor, the resulting mixture has an excess of positive holes. This results from the fact that boron has only three valence electrons, which is not enough to complete the valence bonds. Therefore, when a boron atom displaces a germanium atom, one of the bonds has a hole—that is, a deficiency of one electron. (See fig. 10–2 (C).) The hole in the bond can be filled by an electron from an adjacent bond, but this leaves a hole where the electron came from. Thus, the hole moves and contributes to the conductivity of the substance. Semiconductors with an excess of holes are called *p*-type semiconductors. Aluminum, gallium, indium, boron, and other impurities which make *p*-type semiconductors, are called ACCEPTORS.

As an aid in helping you to remember the two classifications of semiconductors, it should be noted that the word doNor contains the letter N and donor materials are *n*-type; likewise, the word accePtor contains the letter P and acceptor materials are *p*-type.

It should be noted that, generally speaking, electron tubes depend on electrons for conduction. This does not hold true in the case of transistors for they not only make use of electrons but also of holes for obtaining conduction.

THE P-N JUNCTION AS A SEMICONDUCTOR DIODE.—Figure 10–3 shows a *p-n* junction that is made up of an *n*- and a *p*-type semiconductor. When a d-c voltage is applied to the ends of this junction, the amount of current flow that can take place will depend on the polarity of the applied potential. Such a junction is made up of a single piece of material (a single crystal), but the material on each side of the junction has different electrical characteristics.

The construction of a *p-n* junction is an involved process and it is not possible to make a satisfactory junction by taking a piece of *n*-type and *p*-type material and putting them together. Since it is not feasible to construct a satisfactory junction mechanically, two other methods have been developed which are known as the ALLOY or FUSION JUNCTION and the GROWN JUNCTION.

A grown junction is produced by changing the melt (molten germanium or silicon with high order of purity) through the addition of impurities. At the start of crystal growth an impurity of one type is added and near the middle of the growth an opposite type impurity is added. Thus, a junction is formed when the melt is changed from a *p*-type to an *n*-type or from an *n*-type to a *p*-type. A fused junction is produced by placing a small dot of indium on a wafer of *n*-type germanium and causing the two to fuse together through the application of heat. When the fusion takes place, the indium produces *p*-type germanium just below the surface. Thus, there is

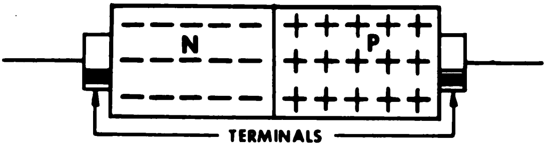


Figure 10–3.—A *p-n* junction.

a *p-n* junction between this region and the remainder of the *n*-type wafer.

In a *p-n* junction should the negative terminal of the power source be connected to the *n*-type material, free electrons will move away from the terminal and toward the junction of the two substances. In a similar fashion, the holes in the *p*-type material will migrate away from the positive terminal towards the junction, where they can combine with the free electrons of the *n*-type material. The junction under these conditions permits a free exchange of positive and negative charges (holes and electrons), and a comparatively large current flow may take place. An applied voltage of this polarity is referred to as forward bias.

If the polarities of the potential applied to the junction are reversed, holes in the *p*-type material are attracted away from the junction to the negative terminal of the applied voltage. Similarly, in the *n*-type the electrons are attracted to the positive terminal until the counter attraction of the charges on the immovable impurity atoms stops the flow. The region close to the junction has all its conduction electrons and holes removed, thereby becoming an insulator. Such a region is called the BARRIER LAYER. This must not be taken to mean that an actual physical barrier exists there but merely a potential barrier. The barrier layer is a temporary insulator only because of the absence of holes or electrons. An applied voltage of this polarity is referred to as a REVERSE BIAS. Under these conditions, a high resistance would result, a relatively small amount of current would flow, and a strong electrostatic field would be developed. This small reverse current is referred to as REVERSE SATURATION CURRENT.

HOLE AND ELECTRON INJECTION.—If a metal point-contact is made to *p*-type material, and if the metal is more negative than the material, the point-contact supplies electrons to the electron-deficient semiconductor. In a similar manner, the *n*-type material has an excess of electrons. A metallic point-contact that is charged positive with respect to the *n*-type material draws off electrons and leaves holes. This is known as HOLE INJECTION.

CHARACTERISTICS OF TRANSISTORS

In the explanation that is to follow, information will be presented that deals with different types of transistors. The junction type transistor is one of the most common used, and since the theories involved in its operation generally apply to the operation of the other types, it will be the only one described in detail.

THE JUNCTION TRANSISTOR.—Keeping in mind the information given in connection with the *p-n* junction, it is easy to see, as illustrated in figure 10–4, that such information can be applied to the junction triode transistor. This transistor is essentially two junctions and is constructed by using three layers of semiconductor material.

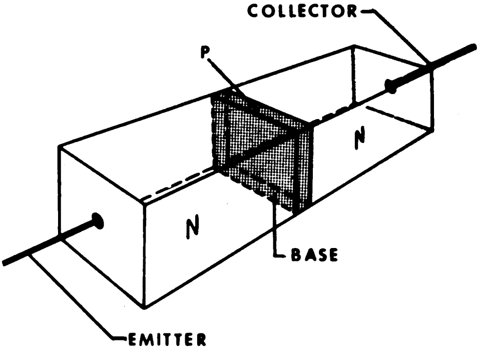


Figure 10–4.—A junction triode transistor.

The center layer is a wafer of either *n*-type or *p*-type material and the outer layers are always of the opposite type. If the center wafer is *p*-type material, the outer layers will be of an *n*-type material and the transistor will be an *n-p-n* junction transistor; this is the type shown in figure 10–4. On the other hand, if the center wafer is *n*-type material, the outer layers will be *p*-type and the transistor will be a *p-n-p* type.

In this type transistor low-resistance connections are made to the outer layers, one of which serves as the EMITTER and the other as the COLLECTOR. A third low-resistance connection to the center wafer provides the BASE connection. The complete assembly is hermetically sealed in a suitable casing to provide ruggedness and freedom from atmospheric contamination.

Such a construction is shown in figure 10–5. The connections to the electrodes are brought through an insulated material to provide external pin connections. This allows the transistor to be plugged in a socket in a manner similar to subminiature vacuum tubes.

In the transistor when the emitter electrode is d-c biased for forward current flow, the emitter effectively emits current carriers into the base region of the transis-

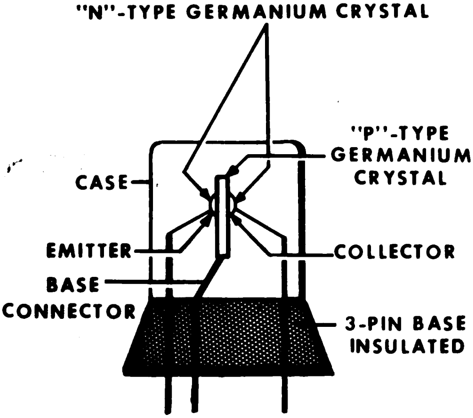


Figure 10–5.—Hermetically sealed junction transistor.

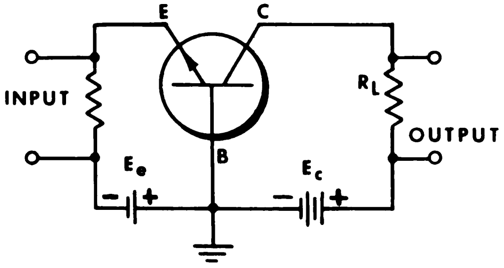
tor. Thus, this electrode is called the emitter. When the collector is d-c biased for reverse current flow, it effectively collects the current carriers, thus it is called the collector electrode. When the emitter is an *n*-type material, the emitted carriers are electrons, while in the *p*-type emitter the emitted carriers are holes. Since this transistor can be looked upon as two junctions, the emitter and base can be regarded as one *p-n* junction and the base and collector as another *p-n* junction.

Figure 10–6 shows the basic circuit connections for the two types of transistors. The transistor in each circuit is represented by the commonly used symbol. The base is represented by a straight line with the emitter and the collector represented by lines at angles to the base. The emitter is the line that contains the arrow. When the arrowhead points toward the base, the transistor is the *p-n-p* type. When the arrowhead points away from the base, the symbol is the *n-p-n* type.

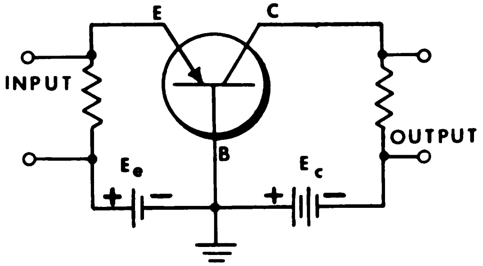
It should be noted that the emitter arrow indicates the direction of positive current flow with forward bias. Positive current flows in the opposite direction from the flow of electrons. This is worthy of attention since many publications will indicate the direction of current as being opposite to the conventional flow of electrons.

There is some tendency to use the same symbol for both types of transistors. Where there is any doubt about the type of transistor indicated on a schematic diagram, the type number of the transistor should be carefully checked prior to replacement or installation.

In figure 10–6 (A) the *p-n* junction between the emitter and the base is biased in the forward (low resistance) direction. The *p-n* junction between the base and the collector is biased in the reverse (high resistance) direction. In other words, in the *n-p-n* transistor the emitter is biased negatively in respect to the base and the collector is biased positively with respect to the base. In the circuit shown in figure 10–6 (B) which uses a *p-n-p* transistor, the electrode potentials are reversed. In figure



NPN
(A)



PNP
(B)

Figure 10–6.—Basic transistor circuits.

10–6 (A) the current flow consisting mainly of electrons will be primarily from emitter to collector. Electrons will leave the emitter and travel into the positive base material. Since the base is very thin, most of the electrons will travel on towards the positive potential on the collector due to the strong electric field existing at the collector junction. The base current will be very slight, consisting only of the few electrons that combine with the holes in the base material. The emitter and collector currents will be approximately equal.

It should be noted that the collector current is controlled mainly by the emitter current rather than the collector voltage. By varying the current flow in the emitter circuit, collector current can also be varied. In practice this variation of emitter current is produced by adding a small a-c signal in series with the d-c bias current. In the circuits shown in figure 10–6, if the emitter voltage is reduced to zero, a small amount of collector current (cutoff or reverse saturation current) will flow through the high resistance of the collector junction. This current is referred to as collector cutoff current, I_{co}.

TRANSISTOR PARAMETERS.—Since the base-emitter circuit of figure 10–6 (A) is biased in its forward direction, it forms a low impedance circuit. The collector (output) circuit is biased in its reverse direction and forms a high impedance circuit. It is this difference between output and input impedances that permits gain to be obtained and the transistor to be used as an amplifier. Even though emitter and collector currents are approximately equal, appreciable power gain can be obtained. In the *p-n-p* junction transistor (fig. 10–6 (B)), power supply voltage polarities are reversed and the major conduction is by holes instead of electrons. Otherwise the operation of the two transistors is similar and the input and output impedances are still low and high, respectively.

It should be remembered that emitter current is practically independent of collector voltage. For example, in a typical grounded-base transistor a 40-volt change in collector voltage will change the emitter current only about 30 percent. It should also be noted that since a small emitter voltage will produce a large emitter current, the emitter circuit will appear as a low resistance to small input signals. Thus, the INTERNAL RESISTANCE of the transistor input circuit will appear very small.

The characteristic curves in figure 10–7 are static voltage and current curves for an *n-p-n* junction transistor in a grounded-base circuit. They are similar to vacuum-tube curves and may be used to show transistor operation. By superimposing loadlines on these curves, as shown in the figure, it is possible to determine proper operating conditions for transistor circuits. These curves

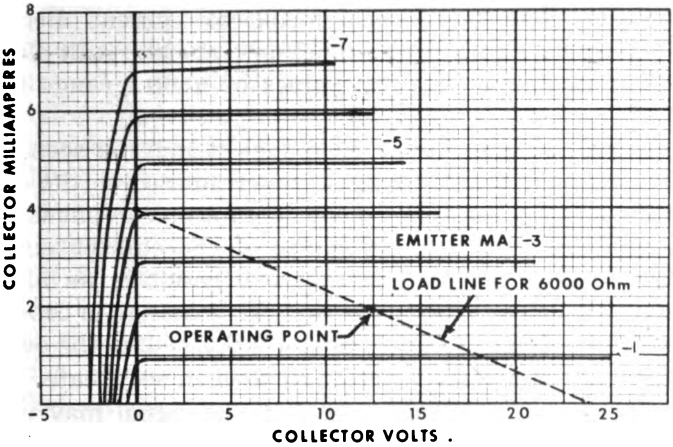


Figure 10–7.—Characteristics of an *n-p-n* junction transistor in a grounded-base circuit.

show collector voltage (v_c) vs. collector current (i_c) for various fixed emitter currents (i_e).

From the characteristic curves shown in figure 10–7 it may be seen that the collector current is practically independent of collector voltage providing the collector voltage is sufficient to collect all the carriers that reach the collector junction.

In the junction type transistor if all the carriers ejected by the emitter reach the collector, the collector current and emitter current would be equal. Under these conditions the emitter-collector current gain would be unity (1). Current gain in transistors is referred to as ALPHA (α) and is analogous to MU (μ) in vacuum tubes.

ALPHA is defined as the ratio of the change in collector current produced by a specific change in emitter current when the collector potential is held constant.

In the *n-p-n* junction transistor a few of the emitted electrons combine with holes of the base material and are drawn to the base terminal. While in the *p-n-p* type a few of the emitted holes combine with electrons of the base material and are likewise drawn to the base terminal. Hence, the increase in emitter current will be slightly greater than the increase in collector current. Because of this fact, ALPHA in the junction transistor will always be less than one. Typical values of ALPHA range from 0.90 to 0.99.

By referring to figure 10–6 (A) it may be seen that when a small signal voltage is superimposed on the emitter bias an appreciable change in emitter current is produced due to the low resistance of the emitter circuit. The change in emitter current will result in an almost equal change in collector current. This change in collector current through the load resistor (R_L) causes the output signal that is developed across R_L to be much greater than the input signal. Since a small input signal may cause approximately the same change in current in the high resistance collector circuit as it causes in the low resistance emitter circuit, amplification is possible.

The following symbols are used in connection with transistors:

Symbol	*Definition*
α	Current gain (amplification factor)
β	Base-current amplification factor
I_b	Base current
I_c	Collector current
I_{co}	Collector cutoff current
I_e	Emitter current
r_b	Base resistance
r_c	Collector resistance
r_e	Emitter resistance
V_b	Base voltage
V_c	Collector voltage
V_e	Emitter voltage
R_i	Input resistance
C_c	Collector capacitance
$f\alpha_{co}$	Alpha cutoff frequency

CORRELATION WITH THE VACUUM TUBE.—In the transistor circuits that have been described the base terminal has been common to the input and output circuits. This arrangement is referred to as the GROUNDED-BASE (COMMON-BASE) circuit. Two other arrangements are possible by making either the emitter or collector the common terminal; these are referred to as either the COMMON-EMITTER or COMMON-COLLECTOR circuits.

Figure 10–8 shows these three circuit arrangements and their triode vacuum-tube equivalents. The grounded-base transistor amplifier, shown in (A), has a comparatively low input impedance (about 50 ohms), a high output impedance (about 100,000 ohms), and no change of signal polarity between input and output. In these respects it is similar to the grounded-grid vacuum-tube amplifier; the emitter corresponds to the cathode, the base to the grid, and the collector to the plate. The grounded-emitter operation as shown in (B) is similar to grounded-cathode tube operation in that it has relatively high input impedance (about 1,000 ohms), high output impedance (about 30,000 ohms), high gain, and a reversal in polarity between input and output. Similarly, as shown in (C), grounded-collector operation corresponds to cathode-follower tube operation in having high input impedance (about 30,000 ohms), low output impedance (about 600 ohms), and no phase reversal.

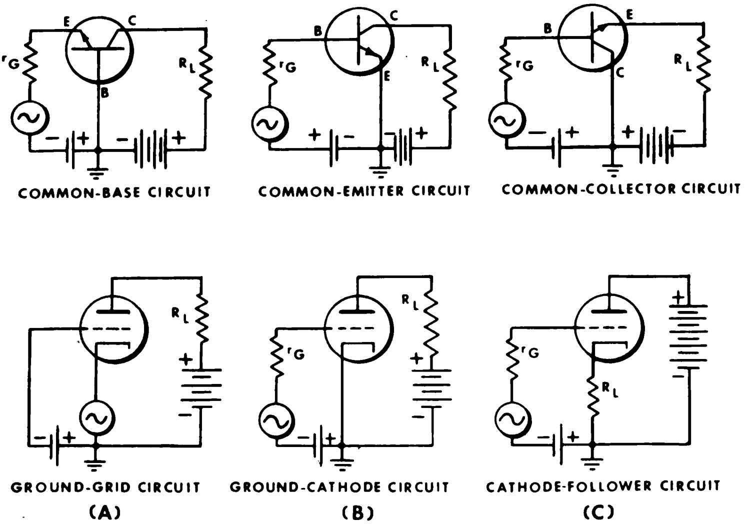


Figure 10–8.—Analogy of transistor and vacuum-tube circuits.

The common-emitter circuit is the only arrangement which produces a phase reversal between output and input signals. The circuit requirements will determine the particular arrangement used and these will be discussed in greater detail later in this chapter. In comparing transistors with vacuum tubes, probably the greatest difference is due to the fact that there is no isolation between the transistor's input and output circuit. It should be remembered that in tube circuits there is considerable isolation between the input and output circuit.

The existence of both *n-p-n* and *p-n-p* types of transistors makes possible a wide variety of applications. This situation is comparable to the imaginary state of affairs that would exist if vacuum tubes were available that would operate with negative plate voltage and positive electrons, in addition to the actual ones with positive plate voltages.

THE POINT-CONTACT TRANSISTOR.—Another type of transistor which functions somewhat similar to the junction transistor is the point-contact. Even though these two types are similar in many respects they are different so far as some electrical characteristics are concerned. The point-contact transistor was discovered earlier than the junction, but presently it is being replaced in most applications by the latter. The information that has been given in connection with the junction transistor will find wide application with the point-contact.

The construction of a point-contact transistor is shown in figure 10–9. In this transistor the emitter and collector connections are made through the pointed ends of very small electrodes. The points of these electrodes, which are spaced only a few thousandths of an inch apart, are connected to the base material. The base material may be either a *p*-type or *n*-type germanium crystal.

During manufacture, when the base material is *n*-type germanium, it is assumed that a small area of *p*-type germanium is created under the electrode contacts. This is illustrated in figure 10–10 (A). If the base material

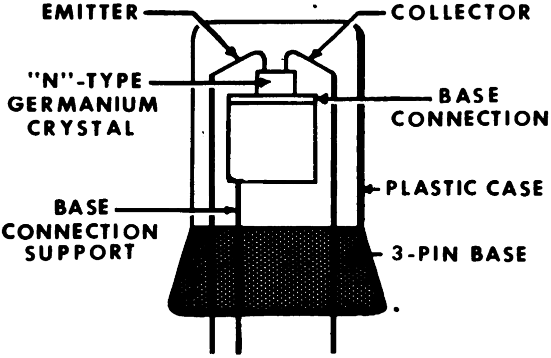


Figure 10–9.—A point-contact transistor.

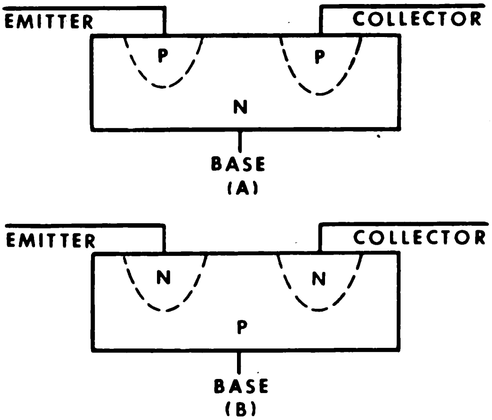


Figure 10–10.—Electrode arrangement of point-contact transistor.

is *p*-type germanium, it is assumed that small *n*-type areas are formed as illustrated in figure 10–10 (B). These assumed areas are created as the result of a forming process in which a current is passed through the electrode during manufacture.

By again referring to figure 10–10 (A) it may be seen that the emitter and collector electrodes make contact with *p*-type material. Thus, the *n*-type point-contact transistor is similar in operation to the *p-n-p* type junction transistor. The same analogy can be made for figure 10–10 (B) as it relates to the *n-p-n* type junction transistor.

The point-contact transistor functions similarly to the junction in that carriers are injected into the base by the forward biased emitter. These carriers are then, as in the junction transistor, collected by the reverse biased collector. One of the major differences between the two types of transistors is that in the point-contact a specific change in emitter current will result in a greater change in collector current. Thus, the current gain (ALPHA) may be greater than one. A typical value of ALPHA is 2.5. Since ALPHA is greater than one in a point-contact transistor, it is possible to have a circuit that exhibits negative resistance. This may be used to an advantage in an oscillator circuit, but may prove undesirable by causing unwanted oscillations in amplifier circuits. An explanation as to why the point-contact transistor possesses a current gain greater than one will not be given since there seems to be no well-founded theory to support such action.

The characteristic curves of a point-contact transistor are shown in figure 10–11. These curves show collector voltage vs. collector current for various fixed emitter currents. By comparing these curves with the ones shown in figure 10–7 for the junction transistor it can be seen that the curves of figure 10–11 have a greater slope. This comparison shows that the transistors possess different characteristics. The point-contact curves show that the collector current is always greater than the emitter current; they also show that there is a substantial increase in collector current when the collector potential is increased.

Another advantage of the point-contact over the junction transistor is its ability to operate at higher frequencies. This is true since the collector capacitance (C_c) is smaller due to the smaller area of the collector junction. Also, since the base resistance (r_b) is lower the effective ALPHA CUTOFF ($f_{\alpha_{co}}$) is higher. ALPHA CUTOFF

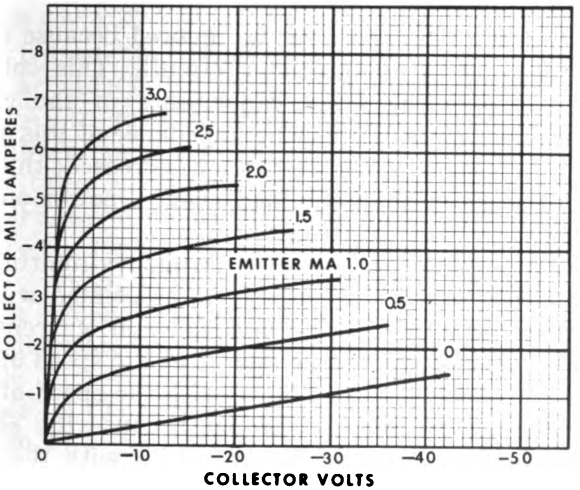


Figure 10–11.—Collector characteristics for point-contact transistor.

will be explained under the next heading, FREQUENCY CHARACTERISTICS.

Some of the newly developed transistors that operate at frequencies higher than the point-contact are the TETRODE JUNCTION, *p-n-i-p* JUNCTION, and SURFACE-BARRIER. Due to the scarcity of information dealing with these newer type transistors extensive coverage will not be given, however, they will receive limited treatment under, FREQUENCY CHARACTERISTICS.

FREQUENCY CHARACTERISTICS.—A limiting factor in the utilization of transistors has been their frequency limitations. Some of the newer types of transistors now possess high-frequency characteristics that make them applicable for many uses in naval electronic equipment. As in the vacuum tube, the high-frequency limitations of transistors are due to the interelectrode capacitances and transit-time effects.

Under forward bias conditions the emitter junction capacitance may be rather large; however, it may be neglected since it is shunted by the low forward resistance. On the other hand, although the collector is biased reversely and causes a rather low collector capacitance

(C_c), 10 to 20 $\mu\mu$f., it cannot be ignored because of the high impedance of the collector circuit. This collector capacitance plus the stray capacitance of the circuit shunts the load impedance. The impedance of this shunting path decreases as the frequency increases; thus, the current change through the load (hence, power output) will drop as frequency increases.

As already mentioned, another frequency limiting factor is caused by transit-time effects associated with the interaction between emitter and collector currents. It should be noted that electron flow is at the speed of light —not because the electrons move with the speed of light (as commonly misunderstood) but because the electric field (E) which, when produced, moves with the speed of light and exerts a force on all the electrons in the material. The hole current, however, depends on the motion of electrons between valence bonds and progresses at a much slower rate. The relatively slow velocity of emitter (hole) current is a contributing factor in frequency limitations.

At the higher frequencies many of the carrier (holes or electrons) injected into the base will fail to reach the collector junction before the input signal reverses its polarity. This occurs because the transit time becomes an appreciable portion of the period of a cycle. Thus, the carriers that fail to reach the collector are trapped in the base material. This results in a decrease in current gain (ALPHA), hence it may be seen that at the higher frequencies ALPHA will decrease as frequency increases.

As previously stated, ALPHA is the ratio of the change in collector current produced by a specific change in emitter current when the collector potential is held constant. This definition applies to the current gain of transistors for low frequency operation, usually 270 c.p.s. ALPHA CUTOFF ($f_{\alpha_{co}}$) is defined as the frequency at which the value of ALPHA drops to 70.7 percent of its low frequency value.

The transit-time effects also result in a phase shift which at ALPHA CUTOFF will be approximately 60 degrees. As in vacuum tubes, these phase shifts are undesirable and reduce circuit efficiency.

Two types of transistors that have been developed that partially overcome collector capacitance and transit time are the TETRODE and *p-n-i-p* (or *n-p-i-n*). The major objective in the development of these transistors was to attain a higher frequency characteristic.

As illustrated by the point-contact transistor, it is possible to reduce collector capacitance (C_c) by reducing the area of the junction. This reduction of area in the junction type transistor is impractical beyond certain limits. It is also possible to reduce collector capacitance by reducing the conductivity of the base material. However, this would increase the base resistance (r_b) which would lower ALPHA CUTOFF.

Figure 10–12 shows a JUNCTION TETRODE TRANSISTOR. It is similar to the junction transistors already discussed except that a second connection is made to the base. This second contact is made to the base material on the opposite side of the normal connection. By applying a repelling bias to the second base connection the emitter-to-collector carrier current through the base material is compressed into a small area near the normal base connection. This action has effectively reduced the cross

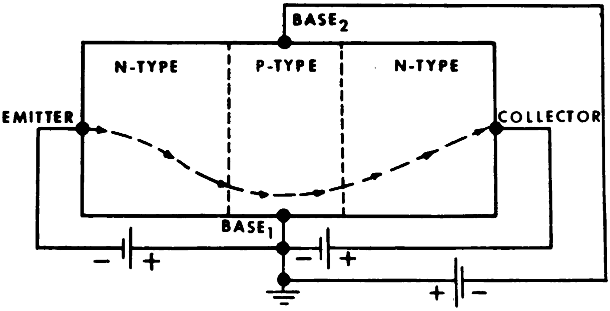


Figure 10–12.—A junction tetrode transistor.

sectional area. Thus, what could not be accomplished mechanically has been accomplished electrically. The major advantage of the TETRODE is that ALPHA CUTOFF can be greatly increased. This is brought about by (1) effectively reducing the cross sectional area of the base material, which reduces collector capacitance, and (2) due to this action it is possible to decrease the thickness of the base material which will reduce the transit time.

An *n-p-i-n* junction transistor is shown in figure 10–13. It is different from the *n-p-n* junction transistor in that an intrinsic layer separates the base from the collector. Intrinsic germanium, referred to as I-type, is either extremely pure germanium, or it has an equal number of donor and acceptor atoms. This layer causes the collector capacitance and base resistance of the transistor to be reduced and thus increases the ALPHA CUTOFF frequency. Even though there is a large voltage drop across the intrinsic material it may be compensated for by increasing the collector voltage.

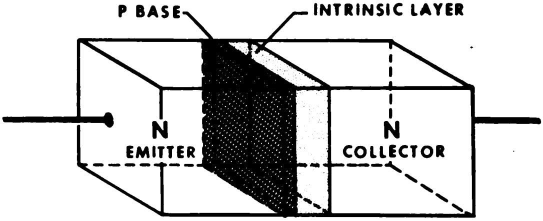


Figure 10–13.—An *n-p-i-n* junction transistor.

POWER.—As stated previously under the heading, GERMANIUM CHARACTERISTICS, thermal agitation at ordinary room temperature causes SOME electrons to acquire sufficient energy to break their covalent bonds and become free. At a certain critical temperature, usually about 200° F. for most samples of germanium, the number of free electrons becomes excessive. The freed electrons from the broken bonds cause the conductivity of even intrinsic germanium to increase sharply, giving germanium a pronounced negative temperature coefficient of resistivity—that is, lower resistance with higher temperature. At the critical temperature, the effect of thermal electrons swamps out the effects of either *n*- or *p*-type impurity conduction, and transistor action disappears.

If the temperature continues to increase, the point is reached at which IMPURITY DIFFUSION takes place. This happens when the thermal agitation becomes so excessive that the impurity atoms will be transferred to the opposite side of the junction. When this happens the transistor will be ruined.

POWER DISSIPATION is important in connection with transistors since the current flow through a transistor, as in any other semiconductor or conductor, produces heat; most of the heat is dissipated by the collector junction. Temperature limitations of transistors will vary with different types and are given by the manufacturer. This temperature limitation should never be exceeded. Various types of high-power transistors have been developed which provide a means of removing the heat from the collector junction. A much used method is to attach one of the transistor electrodes to its container. In most cases the container is connected to the chassis either directly or through a mica insulator which is used to keep the transistor from being grounded. (See fig. 10–14.) The chassis will then serve as a heat sink which aids in heat dissipation and permits a greater power output.

The ambient (room) temperature must be taken into consideration in connection with the power output of transistors. It is easy to understand that the heat dissipation of the transistor will be less when the ambient temperature is high. This is true because a high ambient temperature will decrease the ability of the heat sink to take on heat from the transistor.

A major difference between vacuum tubes and transistors has to do with POWER GAIN. In the transistor

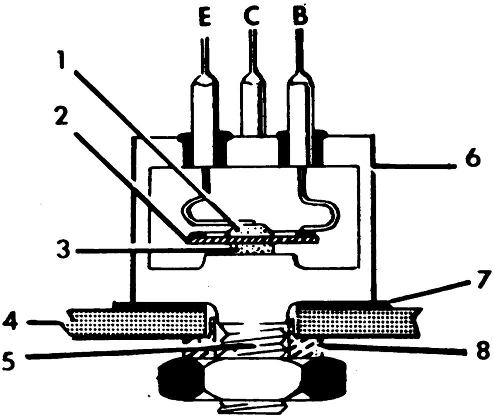


Figure 10–14.—Cutaway view of a typical high-power transistor.

1. Emitter.
2. Base.
3. Collector.
4. Chassis.
5. Mounting stud.
6. Metal case.
7. Mica insulator.
8. Insulator bushing.

amplifier each stage adds to the power gain, while in vacuum-tube amplifiers power amplification is performed only by the last stage. Transistors are similar to vacuum tubes in that, when used as power amplifiers, they also may be connected either single-ended class A, or double-ended (push-pull) class AB or class B. The maximum power gain of transistors is about 40 db. A typical power gain of a junction transistor is 440 (26 db).

NOISE.—The internal noise in transistors is caused by the random variations in the carrier currents when crossing the junctions, and also by the random division of emitter current between the collector and base circuits. The noise factor for a junction transistor is lower than for a point-contact. As compared with vacuum tubes, transistors have a higher noise factor. It is possible to reduce the noise factor by reducing collector voltage. It should also be noted that transistor noise varies inversely with frequency.

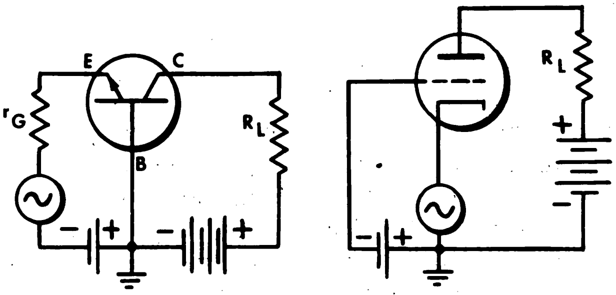
Regardless of similarities in transistors and tubes, as stated previously, their input and output impedances differ radically. A vacuum tube normally operates with its grid-to-cathode circuit biased in the reverse, or high resistance, direction while a transistor's base-to-emitter circuit is biased in the forward, or low resistance, direction. Similarly, in the output circuit, the vacuum-tube plate is biased in the forward direction, the transistor collector in the reverse direction.

The consequence of this is that a vacuum tube has a high input impedance and a low output impedance. The transistor, on the other hand, has a low input impedance and a high output impedance (except the grounded-collector circuit). The resultant circuitry therefore is somewhat different for these two amplifying devices, emphasizing the fact that a transistor is CURRENT CONTROLLED and a vacuum tube is VOLTAGE CONTROLLED. Thus, the transistor operates as a current amplifier and prefers a constant current source. The vacuum tube, on the other hand, operates usually as a voltage amplifier and works best with a constant voltage source.

THE GROUNDED-BASE AMPLIFIER.—The three basic transistor amplifier circuits and their vacuum-tube equivalents were briefly discussed previously. The circuit for the grounded-base amplifier and its vacuum-tube equivalent is shown in figure 10–15. The input signal is applied between the emitter and the base. The bias voltage in the input circuit determines the average emitter potential with reference to the base. The base of the grounded-base amplifier need not be connected to the circuit ground so long as it is common to both input and output circuits.

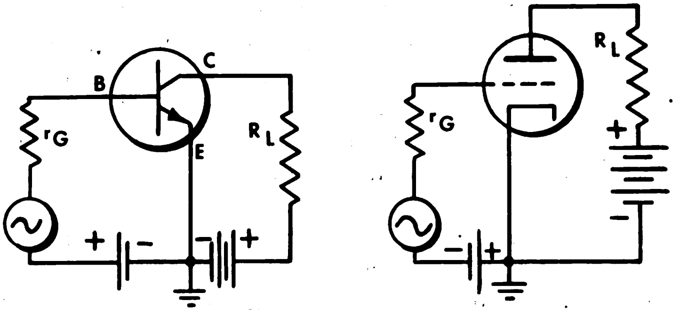
The characteristics of the grounded-base amplifier are LOW INPUT IMPEDANCE, HIGH OUTPUT IMPEDANCE, NO PHASE (POLARITY) REVERSAL BETWEEN INPUT AND OUTPUT, QUITE STABLE, AND MODERATE POWER GAIN.

THE GROUNDED-EMITTER AMPLIFIER.—The grounded-



COMMON-BASE CIRCUIT GROUND-GRID CIRCUIT

Figure 10–15.—Grounded-base amplifier and vacuum-tube equivalent.



COMMON-EMITTER CIRCUIT GROUND-CATHODE CIRCUIT

Figure 10–16.—Grounded-emitter amplifier and vacuum-tube equivalent.

emitter amplifier is shown in figure 10–16. The input signal is applied between the base and emitter. The bias potential in the input circuit determines the average base potential with reference to the common emitter.

The input signal varies the base current about its average value, causing an amplified current variation in the collector circuit. This results in an amplified version of the input signal appearing across the load impedance (R_L).

The characteristic curves of an *n-p-n* junction transistor in a grounded-emitter circuit are shown in figure 10–17. These curves show that the output or collector current (I_c) is much greater than the input or base current (I_b). The reason for this is that the emitter and

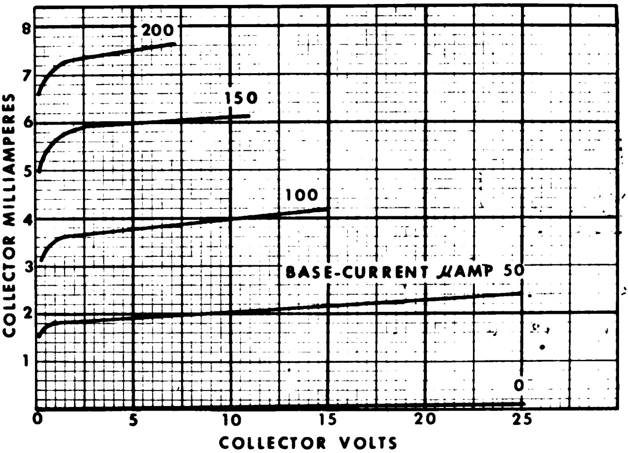
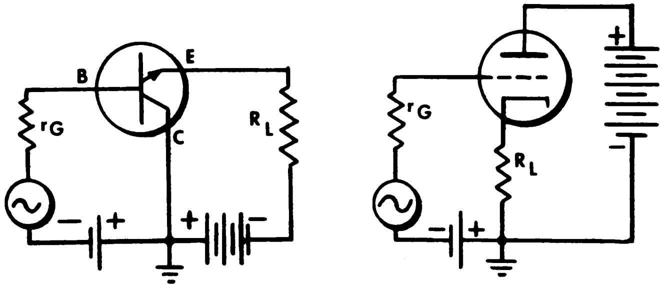


Figure 10–17.—Characteristic of an *n-p-n* junction transistor in a grounded-emitter circuit.

collector currents are almost equal and the base current is only the difference between the currents. This base-collector current amplification, sometimes called BETA (β), is much greater than ALPHA, and expressed in terms of ALPHA is $\beta=\frac{\alpha}{(1-\alpha)}$.

The grounded-emitter is more sensitive to frequency than the grounded-base circuit. This causes the gain to drop to the point that at ALPHA CUTOFF its gain is about equal to that of a grounded-base circuit. The characteristics of the grounded-emitter amplifier are a LOW TO MODERATE INPUT IMPEDANCE, MODERATE TO HIGH OUTPUT IMPEDANCE, THE HIGHEST CURRENT GAIN OF THE THREE BASIC AMPLIFIER CIRCUITS, and a PHASE (POLARITY) REVERSAL BETWEEN INPUT AND OUTPUT.

THE GROUNDED-COLLECTOR AMPLIFIER.—The circuit for the grounded-collector amplifier is shown in figure 10–18. The input signal is applied between the base and the grounded side of the collector circuit, and the output is derived from a common load impedance placed between



COMMON-COLLECTOR CIRCUIT CATHODE-FOLLOWER CIRCUIT

Figure 10—18.—Grounded-collector amplifier and vacuum-tube equivalent.

emitter and ground. Thus, the collector is common to both input and output circuits.

As in the other transistor circuits, bias currents and collector voltages are supplied by d-c sources. It differs from the other basic transistor amplifiers in that the input impedance is relatively high, and the output impedance is very low. Because of this low output impedance, it is often used to match impedances.

The characteristics of the grounded-collector amplifier are HIGH INPUT AND LOW OUTPUT IMPEDANCE, A DEFINITE POWER GAIN (but less than the other two circuits), A VOLTAGE GAIN OF LESS THAN ONE (similar to vacuum-tube cathode follower), and NO PHASE (POLARITY) REVERSAL BETWEEN INPUT AND OUTPUT.

The CIRCUIT GAIN for a transistor is usually specified in decibels of power. These power gains are generally given for each of the three basic types of circuits just described. These values are only typical for the actual circuit gain will depend not only on the characteristic impedances of the transistor but also on how well these impedances are matched to the load and source impedances. The circuit gain also depends on the potentials applied to the electrodes and, as stated previously, the type of basic circuit employed.

TRANSFORMER AND R–C COUPLED AMPLIFIERS.—In order to obtain maximum gain, as has just been explained, it is easy to realize the importance of proper interstage coupling. Transformer coupling is probably the most common and simplest method of interstage coupling used with transistors. Figure 10–19 shows a transformer coupled class A amplifier employing a grounded-emitter circuit.

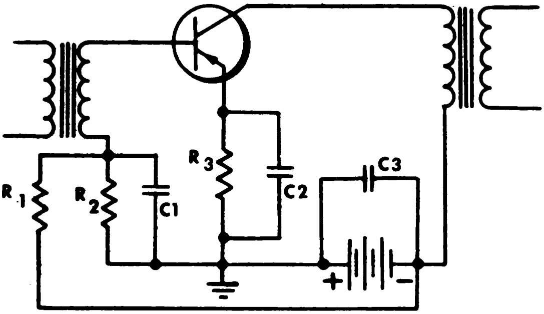


Figure 10–19.—Transformer coupled transistor amplifier.

It should be remembered that the small signal behavior of transistor circuits depend upon the emitter bias current and the collector bias VOLTAGE. Furthermore, the base-to-emitter internal resistance will be different for transistors of the same type. This resistance will also change as ambient temperature changes. Thus it may be seen that by placing a resistor, such as *R*3 in the figure, in series with the emitter, the emitter's current will be rather independent of changes in base-to-emitter resistance. This is true providing the resistance of *R*3 is relatively high in comparison to the low base-to-emitter resistance. Hence, the operation of this circuit will not change considerably with variation in temperature or transistor replacement so far as base-to-emitter resistance is concerned.

The circuit arrangement obtains both emitter current and collector potential from the same battery. This is accomplished by using the divider network *R*1 and *R*2. Resistor *R*2 is made rather small since it is in series with the collector circuit which should have a low resistance. This low resistance enables the collector voltage to be relatively independent of collector current. Capacitors *C*1, *C*2, and *C*3 provide a bypass for the audiofrequencies.

One stage of an R–C coupled transistor amplifier is shown in figure 10–20. Resistors *R*1 and *R*2 provide the base-to-emitter bias voltage in the same manner as described in connection with the transformer coupled stage. Likewise, *R*3 reduces the variation in collector current that may be produced by changes in temperature or transistor replacement. A major difference between the R–C and transformer coupled transistor circuit is that *R*2 in the R–C circuit is not bypassed by a capacitor since this would short out the input signal.

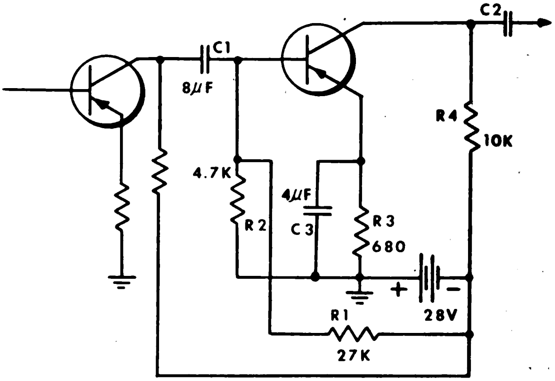


Figure 10–20.—Resistance-capacitance coupled transistor amplifier.

Resistors *R*1 and *R*2 must be large compared to the input resistance. Coupling capacitors *C*1 and *C*2 must be rather large in order to have a reactance at the lowest operating frequency that is low compared to the base-to-emitter circuit resistance. While the low base-to-emitter circuit resistance in transistor circuits require rather large capacitances compared to vacuum-tube circuits, the voltage ratings are low, thus the physical size of the capacitor may still remain small.

CASCADING TRANSISTORS.—By connecting transistors in cascade it is sometimes possible to decrease the equipment's size and weight. However, of the nine possible configurations using two stages, only four are practical. Of these the grounded-emitter to grounded-emitter, illustrated in figure 10–21, is used most and is the only arrangement that will be described. Only junction transistors are used in cascading arrangements for point-contact circuits tend to be unstable.

The principal advantage of the grounded-emitter to grounded-emitter arrangement is that it provides the greatest gain under stable operating conditions. The overall power gain of this circuit arrangement will be approximately 65 db. This circuit is characterized by

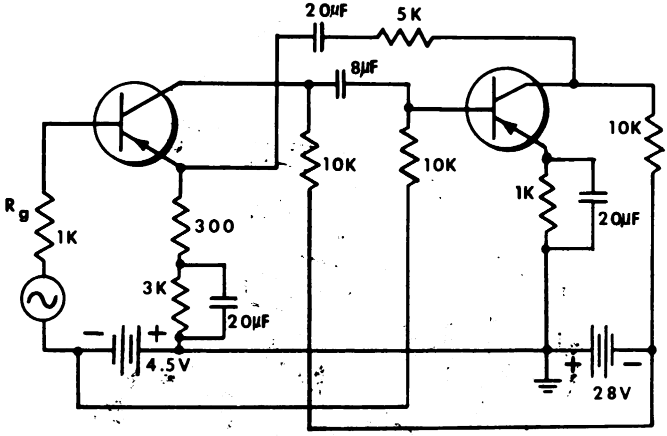


Figure 10–21.—Grounded-emitter to grounded-emitter cascaded transistor amplifier.

medium input and output impedances and is a worthwhile improvement over a single stage. Each grounded-emitter stage provides a phase (polarity) reversal; thus, the input and output voltages are in phase and of the same polarity.

A means of increasing the stability of transistor circuits is through negative feedback. While negative feedback in general decreases the gain of an amplifier, it provides an overall beneficial effect. It permits the amplifier to be stable over a wide range of temperatures and provides a wider frequency response. The circuit in figure 10–21 obtains feedback through the R–C series network that is connected between the emitter of the first stage and the collector of the second stage. The other circuit components operate in much the same manner as those that were described in connection with figure 10–20.

A Typical Transistor Amplifier

An example of transistor equipment development is shown in figure 10–22. This is the interphone amplifier used in some naval aircraft. One of these amplifiers is

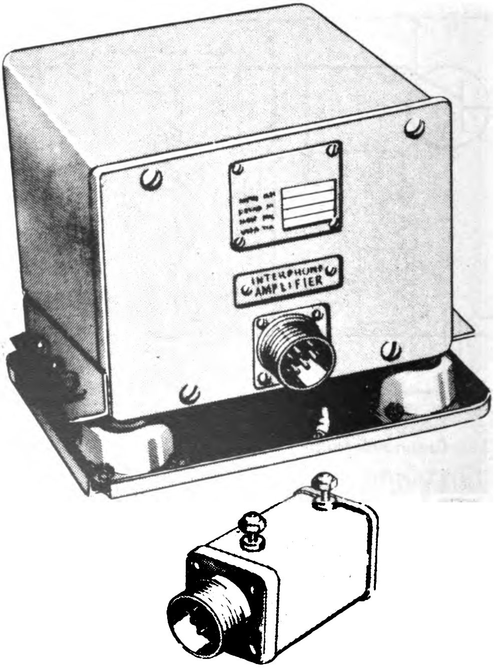


Figure 10–22.—Interphone amplifiers, vacuum-tube and transistor types.

installed at each crew station in the aircraft. Each amplifier has two channels; one boosting the signal from the microphone to the distribution line, the other boosting the signal from the line to the headphones. The large black box in the figure is the vacuum-tube amplifier used in most aircraft. The small box is an experimental transistor amplifier designed to replace the vacuum-tube amplifier.

A schematic of the transistor amplifier is given in figure 10–23. It contains two independent amplifier channels, one for the microphone and the other for the headset. Except for the d-c supply to the carbon microphone in the upper channel, both are the same. All amplifier stages are transformer coupled, to match the input and output impedances of the stages and to keep the d-c operating currents independent of each other. Six CK–

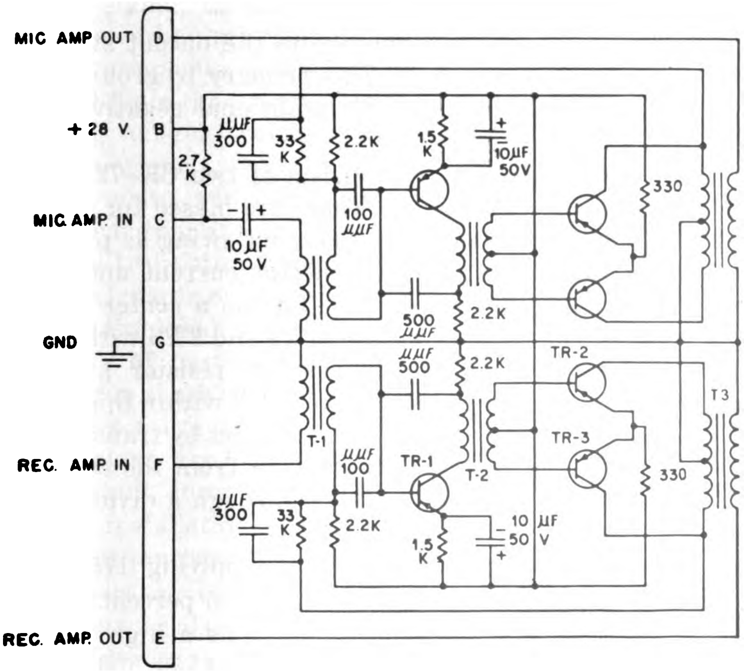


Figure 10–23.—Schematic diagram of transistor amplifier.

721 transistors are used in this unit. These are *p-n-p* junction transistors designed for low power audio and low radiofrequencies.

CIRCUIT ANALYSIS.—For an analysis of the circuit, consider the lower, or receiving, channel on the schematic. Signals from the line come in at pin *F* and enter the primary of *T*1. One end of *T*1's secondary goes directly to the base of *TR*1. The other end goes to the plus 28-volt emitter supply line (signal ground) via a 2.2-K resistor across which a negative feedback signal from the output stage is developed. This resistor in conjunction with the 33-K resistor forms a divider network which provides the base-emitter circuit bias. A 100-μμf. capacitor across *T*1's secondary improves tone and helps to prevent oscillation. Stability for *TR*1 is provided by the 1.5-K emitter resistor which is bypassed by the 10-μf. electrolytic capacitor.

The output of *TR*1 is coupled to the output stage by *T*2. A 2.2-K resistor from *T*2's primary to ground, and the 500-μμf. capacitor, provide additional negative feedback around this first stage.

The power output stage consists of two CK–721 transistors working PUSH-PULL. These are biased for class B operation to permit as large a current swing as possible. As in all class B amplifiers, collector current under no-signal condition is nearly zero. *T*2 has a center tapped secondary to supply the bases of *TR*2 and *TR*3 with push-pull signals. An unbypassed emitter resistor provides some degeneration or negative feedback within this stage. Output signals are coupled to the headset by transformer *T*3. More negative feedback is taken from the collector of *TR*3 around to the input stage through a divider consisting of the 33-K and 2.2-K resistors.

This output stage is capable of supplying 175 milliwatts of audio power at a maximum of 5 percent distortion. This would not exactly qualify as a high fidelity system, but it will drive a loudspeaker at adequate room listening volume and is more than sufficient for headphone operation for which it was designed. Transistors are available today that are capable of delivering 10 watts in a circuit of this type.

A great deal of negative feedback is used in the amplifier, which has several beneficial effects. It permits the unit to be quite stable over a wide range of temperature, from minus 65° F. to plus 160° F. It provides a wider frequency response range than the vacuum-tube unit—from 70 to 20,000 cycles. It also provides a low output impedance, with the output level changing by only 4 or 5 decibels when from 1 to 12 stations are across the line.

ADVANTAGES OF THIS UNIT.—One important advantage of this transistor amplifier unit is that all power is taken from the 28-volt d-c primary power line. Maximum current drain is 42 milliamperes during maximum signal output. Current drain drops to 16 milliamperes at no-signal, 12 milliamperes of this current being used by the carbon microphone. Due to the high sensitivity of the microphone channel, the microphone current was reduced from the usual value of 50 to 12 milliamperes, providing quieter operation and longer life. The amplifier will continue to operate until the 28-volt supply drops below 6 volts, with some reduction in maximum available power output at 6 volts.

The transistor amplifier is quite rugged mechanically. The wiring is on etched circuit boards and the entire unit is encapsulated in casting resin. This eliminates any possibility of parts shaking loose and does away with the need for shock mounts. Since the transistor should have as long a life as any other component, no maintenance provisions are needed. Furthermore, one of the most important advantages is the large saving in weight. In some installations where a number of these units are used as much as 108 pounds may be saved. This is possible because each transistor amplifier weighs only 11 ounces and also no dynamotors are needed since the transistors operate directly from the 28-volt d-c supply.

GAIN CONTROL.—There are various methods of controlling the gain of transistor amplifiers. One method of control is by providing a means of controlling the emitter current of the transistor. When emitter current is reduced the gain will be reduced because the reduction in emitter current will reduce the collector current and thus the amplitude of the signal developed across the output load. This may be accomplished by using a potentiometer to control the base-emitter circuit's bias potential.

RF AMPLIFIER CIRCUITS.—The high-frequency characteristics of transistors, as previously described, are a limiting factor in the application of transistors to radio-frequency amplification. Transistors are being used in RF circuits but special circuit techniques are involved when this is done. For example, in a grounded-emitter to grounded-emitter RF amplifier the collector-to-base capacitance of the transistor necessitates the use of neutralization. This may be accomplished by providing an R–C series of feedback loop from the base of one amplifier stage back to the base of the preceding stage.

In order to provide high gain, grounded-emitter to grounded-emitter circuits are frequently used. The difficulty of providing an impedance match between the high collector impedance of one stage and the low base impedance of the following stage may be overcome by tuning only the primary of the coupling transformer.

TRANSISTORS IN OSCILLATOR CIRCUITS.—In order to design a circuit that will oscillate it is necessary to utilize a device that will provide amplification. A transistor has this ability and, in fact, it will oscillate at frequencies even higher than those at which it may be used as an amplifier. To cause a circuit to oscillate (as in vacuum-tube oscillators), it is necessary to provide either an external feedback path or to make use of the internal feedback of the transistor.

A tickler FEEDBACK TYPE OSCILLATOR is shown in figure 10–24. In this circuit a part of the output current is fed back to the input. Since this a grounded-emitter circuit, the output current is 180 degrees out of phase with the

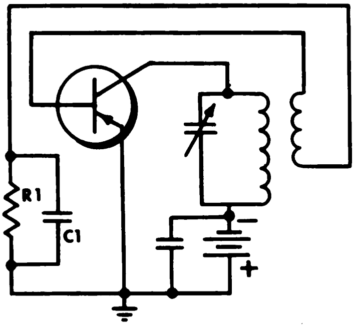


Figure 10–24.—A tickler feedback oscillator.

input. The transformer coupling provides an additional 180-degree phase shift so that the feedback is in phase with the input. Resistor *R*1 and capacitor *C*1 provide self-bias to the base-emitter circuit.

A negative IMPEDANCE OSCILLATOR using a point-contact transistor is shown in figure 10–25. This circuit arrangement is possible because the point-contact transistor has a negative input resistance since its gain is usually greater than 2. It is possible to provide controlled oscillations by shunting the negative input resistance with a tuned circuit. Even though it is necessary to use a point-contact transistor in this circuit, a junction transistor may be used in the external feedback type.

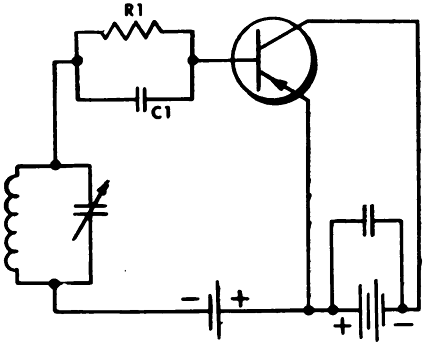


Figure 10–25.—A negative impedance oscillator.

A crystal oscillator such as the one shown in figure 10–26 may be used to provide improved accuracy and stability of frequency. The circuit shown is of the grounded-base type.

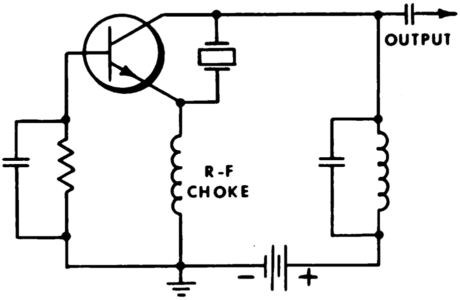


Figure 10–26.—A grounded-base crystal oscillator.

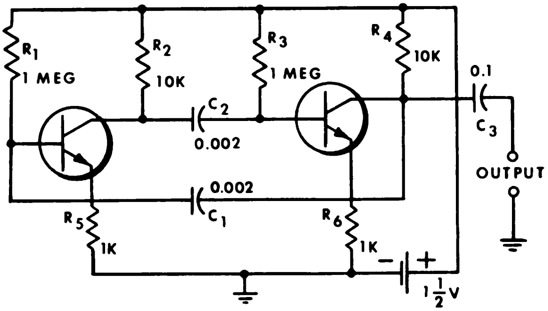


Figure 10–27.—A transistorized multivibrator.

A FREE-RUNNING MULTIVIBRATOR utilizing two transistors is shown in figure 10–27. This circuit is very similar to its vacuum-tube counterpart and is actually a two-stage R–C coupled transistor amplifier. The output of the second stage is fed back to the input of the first stage through capacitor *C*1. This circuit will produce a rectangular output waveform.

The Servicing and Care of Transistors

LIFE EXPECTANCY.—Transistors, due to their simplicity and ruggedness, have a very long life. One of their outstanding features is their practically indefinite life expectancy. Long life was originally predicted on the basis of transistor construction and conduction which indicated that there was nothing that can wear out. Available data indicates that this prediction was true with most units. The most common type of failure comes about gradually after a long period of time and several thousand hours of operation. Failure here is due to the gradual rise of saturation current (collector current when the base current is zero). While various self-bias and fixed-bias circuits can minimize this, efficiency and gain will suffer. In an amplifier circuit, this effect decreases the available volume. Gradually, as the limit of the automatic biasing arrangement is reached, there is a noticeable increase in distortion.

Another type of failure occurs in some circuits because of HIGH TEMPERATURES. As temperature increases, the quiescent current increases, causing the heating effect on the collector junction to increase. This, in turn, increases quiescent current, and so on, until runaway current destroys the transistor. A large resistance in series with the emitter is often used to prevent this condition, similar to the cathode bias resistor of a vacuum tube.

Although transistors do not require a vacuum, they must be HERMETICALLY SEALED. Water vapor will quickly contaminate any unprotected junction. It has been found that the saturation current may increase to 100 times its dry air value when the relative humidity rises to 50 percent. New USN specifications require all transistors to have hermetically sealed metal cases, since plastic cannot be relied upon to remain moistureproof.

MAINTENANCE TECHNIQUES.—The most heartening fact about transistor equipment is that maintenance will not be required as often as for vacuum-tube gear. However, transistor equipment is extremely compact and will gradually become even more so and this will complicate maintenance.

The 100-watt soldering iron and water pump pliers often found on electronic work benches are somewhat clumsy for transistor work. Most pliers, including needle nose types, are too large; however, the smaller ones can be ground down to a manageable size. Tweezers are helpful in connection with transistor work and may be used in place of pliers in many instances. The common types of diagonal cutters are usually too broad but can often be modified by grinding away some of the metal on the outside of the jaws.

Transistors are VULNERABLE TO OVERHEATING, and soldering must be performed with care. Like many subminiature tubes, the transistor pigtails are usually soldered into the circuit. A soldering pencil with a long thin tip is required. In addition clip-on type heat shunts are needed. These shunts (fig. 10–28) are heat dissipating devices that are clamped across the pigtail lead between the transistor and the tie point terminal to be soldered. The heat which is applied to the terminal by the soldering iron, and which would normally flow directly to the transistor and damage it, is effectively shunted to the heat shunt clamp and dissipated. The clamp should be allowed to remain for 15 to 20 seconds before removal. These clamps should be used on the leads of all subminiature components, to keep the heat of soldering away from the delicate transistors and electrolytic capacitors. The solder used should be 60/40 since it has the lowest melting temperature of all common solders; it will be most conveniently used in the form of 1/16-inch diameter multicore wire solder.

Good work-area lighting is necessary, preferably in the form of a movable spotlight, and a large magnifying lens will be quite welcome. The latter should be equipped with a holder so that both hands are free.

Ratings of each component must be checked before any tests are made. In low-impedance, current-operated transistor circuits, many capacitors are electrolytics of very low voltage rating. A typical coupling capacitor will have a capacity of 3 μf., a maximum voltage of 3 or per-

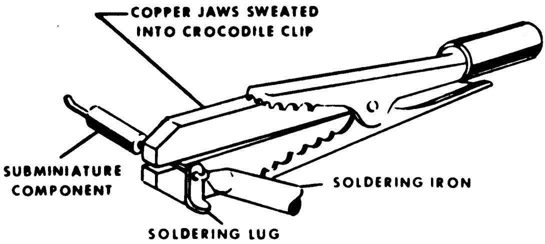


Figure 10–28.—A heat shunt.

haps 6 volts, and will be about the size of a No. 8 machine screw. One application of a 22½-volt battery operated ohmmeter or a capacity bridge which applies 50 volts, and the capacitor may be ruined. As with all electrolytics, the polarity of the test voltage is as important as the magnitude. New electrolytics will often require forming before they will work in a circuit. Connect them to a battery of their rated voltage in series with a 10,000-ohm resistor until they stop drawing current.

Always ascertain whether a transistor is *n-p-n* or *p-n-p* before applying voltages. Vacuum tubes always take positive plate voltages, and will not be damaged by negative plate voltage. A transistor circuit, however, may include both types of transistors. An inadvertent application of reverse voltage to a transistor may damage it if the maximum power dissipation ratings of the junctions are exceeded.

Transistors are susceptible to transient current surges, such as those caused by sudden signal changes or by disconnection of the transistor. It is always advisable to deenergize the circuit before connecting or disconnecting a transistor. If the transistor must be inserted in a live circuit, always connect the base lead first. When disconnecting, always disconnect the base lead first.

The LOW SIGNAL LEVELS at which transistor circuits operate make stray signal pickup troublesome. Careful bypassing of test leads is often needed to keep out signals such as those produced by local transmitters and 60-cycle hum. Many pieces of transistor test equipment are battery operated to eliminate hum and ground current. It is often helpful to introduce test signals at various stages by means of inductive coupling. A coil is connected across the ends of the signal generator leads, and is lined up with a transformer or coil in the equipment under test.

Extreme caution must be used when CHECKING VOLTAGES OR WAVEFORMS. This is especially important in transistorized equipment due to the compactness of the components. The closely spaced terminals may be accidentally connected together by the test prods which may result in surge currents that can ruin the transistor.

SENSITIVITY TO HEAT is a common trouble with transistor circuits, making a heat lamp and thermometer useful additions to the test equipment supply. A 60-watt light bulb in a box could serve as an oven substitute. Before approving a piece of equipment as being in good repair, operate it over the full range of temperatures given in the equipment specification. Many stubborn cases of oscillation will occur only when transistors are warm, and will disappear after cooling.

QUIZ

1. Transistors are extremely useful in airborne electronics equipment because
   a. transistor equipment is heavy and will not need to be strongly secured
   b. 28 volts for standby operation of transistors is readily available
   c. transistors require little current and no dynamotor supply is needed
   d. transistor equipment is short lived and can be replaced very easily

2. The valance bonds in a semiconductor material are weakened by a deficiency of electrons. This is accomplished by mixing germanium with a/an
   a. donor impurity
   b. acceptor impurity
   c. *n*-type material
   d. *p*-type material

3. Applying reverse bias to a *p-n* junction causes the formation of a strong electrostatic field at the junction. This field is established because of
   a. a large amount of current flow
   b. the low resistance of the junction
   c. the buildup of the barrier layer or temporary insulating region
   d. the large amount of current carriers present in the barrier layer

4. Intrinsic (pure) germanium tends to act as a/an
   a. insulating material
   b. good conductor
   c. *p*-type material
   d. *n*-type material

5. With forward bias applied to a *p-n* junction
   a. no movement of holes or electrons will take place
   b. a free interchange of holes and electrons will take place
   c. the negative terminal is connected to the *p*-type material
   d. the electrons and holes will move way from the junction

6. In the transistor circuit shown in figure 10–16, which type of transistor is depicted?
   a. *n*-type
   b. *p-n-p*
   c. *p-n-i-p*
   d. *n-p-n*

7. Referring to the circuit in question 6, what characteristic does the stage possess?
   a. High input impedance
   b. Low output impedance
   c. Highest power gain of three basic circuits
   d. Lowest power gain of the three basic circuits

8. The triode junction transistor consists of
   a. three dissimilar types of semiconductor materials
   b. a pair of *p-n* junctions
   c. three similar types of semiconductor materials
   d. a pair of *p-n* junctions and two barrier layers

9. A semiconductor material has
   a. better conductivity than a good conductor
   b. the same conductivity as a good conductor
   c. poorer conductivity than a good insulator
   d. better conductivity than a good insulator

10. It is desired to substitute a *p-n-p* transistor in place of an *n-p-n* transistor. In order to accomplish this, you must
    a. reverse the bias polarities
    b. apply reverse bias between emitter and base
    c. apply forward bias between collector and base
    d. increase the bias potentials

11. In a transistor, the base corresponds to what element in a vacuum tube?
    a. The heater
    b. The plate
    c. The control grid
    d. The cathode

12. Voltage amplification is possible in a junction transistor because
    a. more current flows in the collector circuit than in the emitter circuit
    b. the emitter circuit has low impedance and the collector circuit has high impedance
    c. base current is greater than collector current
    d. base current is greater than emitter current

13. α (alpha) is the symbol for
    a. current amplification factor
    b. voltage amplification factor
    c. collector current
    d. collector cutoff current

14. The common-collector circuit for transistors is analogous to the
    a. grounded-cathode vacuum-tube circuit
    b. grounded-grid vacuum-tube circuit
    c. cathode-follower vacuum-tube circuit
    d. common-heater vacuum-tube circuit

15. The *p*-type point contact transistor is similar in operation to the
    a. *p-n-p* junction transistor
    b. *n-p-n* junction transistor
    c. germanium diode
    d. silicon diode

16. The change in collector current of a point contact transistor is greater than the change in emitter current. This is an indication of
    a. alpha less than unity
    b. base current larger than emitter current
    c. unsuitable characteristics for oscillator applications
    d. alpha greater than unity

17. A junction tetrode transistor can be developed by
    a. making two connections to the base
    b. making two connections to the emitter
    c. making two connections to the collector
    d. inserting two bases

18. Alpha cutoff is defined as that frequency where the value of alpha has decreased to 70.7% of its
    a. mid-frequency value
    b. low-frequency value
    c. high-frequency value
    d. cutoff value

19. The junction tetrode transistor gives improved high-frequency response because the
    a. base area has been mechanically increased
    b. transit time has been increased
    c. base area has been electrically decreased
    d. transit time has been decreased by increasing base thickness

20. The junction transistor is basically a/an
  a. voltage controlled device
  b. noise-free amplifying device
  c. power amplifying device
  d. unstable amplifying device

21. In R–C coupling transistor stages, the coupling capacitor must usually be electrically large due to the
  a. high value of collector to base impedance
  b. low value of input resistance in most stages
  c. high value of load resistance in the collector circuit
  d. low value of load resistance in the collector circuit

22. Negative feedback is sometimes used in transistor circuits in order to
  a. decrease stability
  b. increase gain
  c. improve frequency response
  d. reduce the current requirements

23. Transformer coupling is used in transistor circuits mainly for
  a. impedance matching purposes
  b. reducing hum pickup
  c. improving frequency response by providing a more uniform load
  d. increasing power gain due to the step-up turns ratio

24. A heat shunt is necessary when soldering transistors into circuits in order to
  a. dissipate heat before reaching the transistor
  b. heat the connectors to a high enough temperature to melt solder
  c. protect the body of the technician
  d. enable the technician to use a smaller soldering iron

25. A heat sink is a device for
  a. checking transistor operation throughout its specified temperature range
  b. maintaining a constant operating temperature for the transistor
  c. increasing the maximum temperature rating for a transistor
  d. increasing the power handling capabilities of a transistor

CHAPTER 11

SPECIAL MAINTENANCE TECHNIQUES

PREVENTIVE MAINTENANCE involves periodic inspection, overall cleaning and lubrication, checking brushes, cleaning and tightening contactors, adjusting and calibrating electronic circuits, and checking system performance. This maintenance permits the detection of wear, weakness, and faulty adjustment before serious trouble develops. Recommended preventive maintenance operations are given in detail in the *Handbook of Service Instructions* (HSI) for each specific equipment or system. These operations are generally performed in connection with periodic inspections and are listed on the check sheet. Periodic checks will be discussed in detail later in this chapter.

When preventive maintenance tests show that an equipment operates inefficiently or needs repairs or modifications, mechanical and electrical adjustments or the replacement of parts may be necessary. Troubles in electronic systems are found by a process of localization, which consists of system troubleshooting and unit troubleshooting. System troubleshooting involves an understanding of the organization of the major system which may consist of several interrelated minor systems. Thus, the technician should have a good understanding of the overall system in order to perform corrective actions with the least delay and the greatest assurance.

In connection with this work a thorough knowledge of the contents of various publications will prove helpful.

Some of these are:

1. *Handbook of Service Instructions* for the equipments.
2. *Handbook of Maintenance Instructions* for the aircraft.
3. *Handbook of Overhaul Instructions* for the equipments.
4. *Illustrated Parts Breakdowns* for the equipments.

The last three of these manuals have been covered briefly in chapter 1 under Sources of Information; the *Handbook of Service Instructions* will now be covered in detail.

HANDBOOK OF SERVICE INSTRUCTIONS

These service handbooks are prepared for use by organizational and field maintenance personnel as an aid in understanding and performing installation, maintenance, and service procedures for each type equipment. Repair, replacements, adjustments, and recalibration data for the electronic equipment are limited to that capable of being performed by organizational and field maintenance activities. This maintenance may be performed with the common tools, test equipment, and spare parts authorized in allowance lists and those specialized tools and test equipment listed in Section II of these handbooks. Further information concerning any detail part of the equipment may be found in the *Illustrated Parts Breakdowns* for the equipment.

These handbooks provide an excellent source of readily available material for each equipment. It is advisable for the technician to spend some of his spare time in becoming familiar with its contents prior to his having to work on the equipment rather than waiting until the need is at hand. This is especially true of the sections on Theory of Operation and Maintenance Procedures.

Most service handbooks for electronic equipment are divided into eight major sections as follows:

SECTION I—DESCRIPTION AND LEADING PARTICULARS.—This section contains the following information:

1. A brief account of the equipment's purpose, limitations, and how, in general, it accomplishes its mission.
2. Condensed factual data to familiarize maintenance personnel with the equipment's characteristics and physical makeup, such as tube and fuse complement, frequency ranges, input and output characteristics, power requirements, and operating function of controls.

SECTION II—TEST EQUIPMENT AND TOOLS.—This section lists the tools and test equipments required for the disassembly, assembly, and testing of the equipment or system.

SECTION III—PREPARATION FOR USE AND RESHIPMENT.—This section describes and illustrates the following procedures:

1. Unpacking and inspecting the equipment.
2. Preinstallation bench testing.
3. Installation of the equipment.
4. Packing for reshipment.

SECTION IV—THEORY OF OPERATION.—This section presents information that deals with the equipment's theory of operation from the standpoint of maintenance It should prove most helpful to the technician since it presents the functional relationships among the various stages of the equipment and the job that each stage must perform. The theory and circuitry of each stage are explained in detail and in many instances simplified schematics and block diagrams are used.

SECTION V—ORGANIZATIONAL MAINTENANCE.—This section provides instructions essential for the maintenance of the equipment by organizational (squadron) personnel. This maintenance consists normally of inspection, cleaning, servicing, preservation, lubrication and adjustment as required, and may also consist of minor parts replacement not requiring highly technical skills.

SECTION VI—FIELD (FASRON) MAINTENANCE.—This section provides instructions essential for the maintenance of the equipment by field (FASRon) personnel. Thus, it will normally be limited to maintenance consisting of replacement of unserviceable parts, subassemblies, or assemblies, and the alinement, testing, and adjustment of the equipment.

SECTION VII—DIAGRAMS.—This section includes diagrams and associated data necessary for maintenance technicians to:

1. Understand the makeup and function of each intercomponent power, control, and signal circuit.
2. Trace each such circuit.
3. Make continuity checks.
4. Accomplish general and specific trouble analysis on inoperative or malfunctioning circuits.

SECTION VIII—DIFFERENCE DATA SHEETS.—This section provides information to be used with new models of the original equipment or system which reflect minor changes from the original design. A separate difference data sheet(s) is included for each additional model which briefly specifies the differences in service procedures from those covering the original model in preceding sections of the handbook.

ORGANIZATIONAL AND OPERATIONAL MAINTENANCE

This is the level of maintenance that is performed by operational squadrons that are not self-supporting. It consists normally of inspection, cleaning, servicing, preservation, lubrication, and adjustment as required, and may also consist of minor parts replacement not requiring highly technical skill.

Excellent sources of information that deal with periodic inspections are the *Handbooks of Aircraft Inspection Requirements*. These handbooks are written by the manufacturer and are available for most aircraft.

These handbooks are divided into six sections. The first four sections contain basic requirements comprising preflight, postflight, intermediate, and major inspections. Section V contains inspection requirements that supplement the requirements of one of the four inspections at specific periods or upon occurrence of specific conditions. Section VI contains requirements for replacing specific accessories or components at prescribed periods. In order to arrange inspection and replacement requirements as nearly as possible according to the manner in which work will be divided and assigned, the requirements in each section of the handbook are divided into groups under "sytem" headings. A system title indicates either a functional system or a group of related components, such as electrical, structural, radio and radar, etc.

PREFLIGHT INSPECTION.—This inspection is performed prior to the first flight of the day. The inspection consists of checking the aircraft for flight preparedness by performing visual examination and operational tests to discover defects and maladjustments that, if not corrected, could cause accidents or aborted missions.

POSTFLIGHT INSPECTION.—This inspection is performed after the last flight of the day or more frequently when warranted by local conditions. This inspection is, basically, a combination of requirements for checking equipment that requires daily verification of satisfactory functioning, plus requirements that prescribe searching for defects that become apparent after the aircraft is flown. It is intended that evidence of chafing, leaks, and similar conditions be discovered and corrected during the postflight inspection to preclude progression of such a relatively minor problem to a state that would require major maintenance to remedy the deficiency. The postflight inspection is, therefore, an important function that should be performed with care.

INTERMEDIATE INSPECTION.—The intermediate inspection is, basically, a limited overall examination of the condition of the aircraft. The inspection includes certain requirements that are also applicable to preflight or postflight inspections, and it also includes requirements that must be fulfilled at periods occurring more frequently than major inspections. The statement of each requirement indicates how thorough the examination or test should be, and refers to any special tool or test equipment that is to be used. These inspections are usually performed at the following intervals:

1. The first intermediate inspection will normally become due at the expiration of 30 flying hours after the preceding major inspection.
2. The second intermediate inspection will normally become due at the expiration of 30 hours after the first intermediate inspection, normally 60 hours after the preceding major inspection.
3. The third intermediate inspection will normally become due at the expiration of 30 flying hours after the second intermediate inspection, normally 90 hours after the preceding major inspection.

A CALENDAR intermediate inspection is performed in lieu of the first, second, or third intermediate inspection, when the aircraft has not accumulated 30 flying hours in the 60 day period since the last previous intermediate or major inspection.

MAJOR INSPECTION.—The major inspection is a thorough and searching inspection of the entire aircraft. The inspection includes certain requirements that are also applicable to preflight, postflight, and intermediate inspection. The thoroughness of the check required and special tools or test equipment to be used are indicated in the statement of each requirement. The major inspection will normally be performed at the expiration of 30 flying hours after the third intermediate inspection following the preceding major inspection, normally 120 flying hours after the preceding major inspection. A CALENDAR major inspection is performed in lieu of a regular major inspection when the aircraft does not accumulate 120 flying hours in the 180 day period since the previous major inspection.

It is possible to determine if an equipment is functioning properly by consulting the *Handbook of Service Instructions*. Section V contains the MINIMUM PERFORMANCE STANDARDS for the equipment. These standards give indications, readings, and tests which will indicate that a component, equipment, or system will provide minimum acceptable results. These observations will be obtained from indications of equipment characteristics, such as overall receiver sensitivity, frequency accuracy, frequency tracking, power output, radar indicator presentations, accuracy of computer test problems, and other peculiar characteristics or functions. This will aid in establishing that the equipment is functioning properly, or in case of an improper indication, it will localize the trouble to a component assembly or subassembly.

System Trouble Analysis

The technician may be aided in the process of isolating a trouble to a particular equipment component by consulting Section V of the *Handbook of Service Instructions*. This section contains system analysis and troubleshooting procedures and charts. By following these step-by-step procedures it is possible to quickly isolate the trouble to a faulty equipment component, or group of components, of the electronic system.

This section also describes minor repairs and adjustments, such as replacing fuses, pilot lamps, tubes, and such other work which is within the capabilities of operational maintenance personnel. It may also contain a step-by-step procedure of any required alinements or adjustments.

This is the type of maintenance described in Section VI of the *Handbook of Service Instructions*, and is the class C maintenance performed by a Fleet Aircraft Service Squadron or an aviation electronics shop aboard a ship. This maintenance of the equipment is based on a complete test procedure, designed to completely check an equipment component, and trouble isolation procedures, designed to isolate trouble to a detailed part within a subassembly, or to a single circuit within the equipment. It is absolutely essential that the test equipment be properly calibrated, properly terminated, and otherwise in excellent condition.

The minimum performance standards given in Section VI of the *Handbook of Service Instructions* differ from those given in connection with operational maintenance. They provide a more comprehensive and detailed check, utilizing test equipment authorized for use by field maintenance personnel.

TROUBLE ISOLATION.—Efficient troubleshooting calls for an orderly, systematic plan of attack. No use leaping from one end of the equipment to the other end, guessing that "maybe the trouble is here." Troubleshooting at the field level requires a good understanding of the theory of operation of the equipment. A visual inspection should first be made. This will frequently pinpoint the trouble. When making this inspection, look for discolored or burned resistors, discolored or burned terminal boards, corroded switch contacts, broken or frayed wires, gassy tubes, leaking electrolytic capacitors, and loose mechanical assemblies.

The *Handbook of Service Instructions* contains many circuit breakdown illustrations, tube socket voltage and resistance charts, and test bench procedures which will prove helpful to the technician. After making repairs, it is frequently necessary to aline and adjust the component or equipment in order to assure correct performance.

REMOVAL AND REPLACEMENT OF COMPONENTS.—Extreme care should be exercised in the replacement of defective components: make sure that the equipment is deenergized; do not disturb critical adjustments; avoid the rearrangement of wiring as much as possible; if the wiring must be disturbed, try to replace it to its original position; when using a soldering iron, employ a minimum of solder to avoid dropping hot solder on the wiring; make sure that all solder joints are firm.

Close reference should be made to the instructions given in the *Handbook of Service Instructions* when removing or replacing defective components. This is necessary since many of the components, such as cathode-ray tubes, require special procedures. The DETAILED INSTRUCTIONS SHOULD BE FOLLOWED CLOSELY in the removal and replacement of equipment.

ENVIRONMENTAL CONSIDERATIONS

In recent years the effects of environmental conditions upon the operation of electronic and electrical equipment has greatly increased the maintenance problems of the technician. These peculiar conditions may be grouped under the major headings of altitude, temperature, and humidity. At the extremes of these conditions special maintenance and operating procedures are required. Equipments required to function at these extremes frequently fail due to the effect of decreased air density, radical temperature changes, and moisture.

Continuous damp, warm air causes condensation of atmospheric moisture within equipment unless units are hermetically sealed or the interiors are maintained at a temperature higher than surrounding atmospheric temperatures. Condensed moisture forms leakage paths and causes corrosion. These climatic conditions promote rapid fungus growth which in itself has a corrosive effect on materials, such as wire, switch contacts, and other metal parts.

ADVERSE CLIMATIC CONDITIONS AND THEIR EFFECTS.—**HUMIDITY** is a term describing the amount of water vapor in the air. It is usually expressed as a percentage of the total amount of water the air can hold at a given temperature. Thus, 50 percent means the air contains one-half the total water it can hold, and 100 percent means it contains all it is capable of holding. Air can hold more water as its temperature increases. In tropical areas the humidity varies between 60 and 100 percent. This high humidity accounts for the condensation of moisture, or sweating, on various parts of electronic equipments when they undergo temperature changes. Condensed moisture on insulating materials reduces their insulating qualities and results in arc-overs and shorts between terminals. The water vapor also penetrates into the body of insulation, is absorbed, and causes similar effects. High humidity also causes corrosion of metals. Other sources of moisture which cause deterioration include fog, salt spray, and rain.

In general, equipment may encounter extreme **TEMPERATURES** ranging from minus 65° F. to a maximum of 135° F. under various climatic conditions of high humidity, fog, rain, salt spray, salt air, cold, insects, fungi, and dust. High temperature and moisture vapor cause rapid corrosion. Fungus and bacterial growths produce acids and other products which speed corrosion, etching of surfaces, and oxidation. This interferes with the operation of moving parts, screws, and so forth, and causes dust between terminals, capacitor plates, and other parts, which produces noise, loss in sensitivity, and arc-overs.

Variations of temperature cause moisture to be breathed through any small cracks, pinholes, or vents in the equipment. As the temperature rises, the air inside a piece of equipment expands and it is expelled, in part, through the openings and vents. When the temperature falls, the air inside the equipment contracts and outside air is admitted through all openings and vents. The moisture which is breathed destroys the insulating qualities of dielectrics and corrodes metals.

FUNGUS is a form of plant life which feeds on materials of vegetable and animal origin including paper, cotton, and so forth, and such things as dead insects and other fungi. These may be spread by wind, dust, dirt, and insects, such as ants, flies, and mites. Growth may take place on materials other than those of organic origin if a spot of dust or other nutrient substance is present. Fungi thrive in the high humidities and temperatures. Fungus growth causes decay, accelerates the deterioration of insulating materials, and short circuits items, such as relays, jacks, and keys. The inclusion of a fungicidal compound in the manufacture of the equipment retards the growth of these fungi.

CLIMATIC DETERIORATION PREVENTION. — Most new equipment is given a climatic deterioration prevention treatment which provides a reasonable degree of protection against fungus growth, moisture, corrosion, salt spray, insects, cold, desert heat, and so forth. The treatment involves the use of a lacquer or varnish coating material applied with a spray gun or brush. Detailed instructions dealing with this treatment may be found in the *General Instructions Handbook,* T. O. No. 16–1–41.

Special Maintenance Problems

As the ceiling for aircraft on extended flights has been raised higher and higher, many new types of operating and maintenance problems have developed. Some of the various types of equipment involved are generators, voltage regulators, electric motors and solenoids, and radio and radar receivers and transmitters. ELECTRIC BRUSHES on generators, alternators, inverters, electric motors, and other rotating electrical machinery wear away very rapidly at high altitude (around 40,000 feet). Special brushes have been developed that have longer life at high altitude. Thus, it is advisable to use the proper type brushes and to check them more frequently when high altitude flying is being performed.

While most SWITCHES will break a circuit safely at sea level, their contacts may burn and in some cases even melt at high altitude. It has been found that double-break contact switches somewhat alleviate this fault. Since electronic systems use special design switches for high altitude operation, when a replacement is necessary use the proper type.

Other items that often fail during high altitude flights are electrical PLUGS AND RECEPTACLES. A voltage breakdown occurs between the pins and shell along the surface of the insulating material. The result is a burned plug. This happens because the breakdown voltage is less at high altitude. For example, the breakdown voltage for a ¼-inch air gap is about 3.7 times greater at sea level than at 40,000 feet.

This condition may be overcome by sealing the connector with a potting compound. This reduces the probability of arc-over between pins or between pins and the shell of the electric connector since the dielectric characteristics of the connector are improved. This sealing compound will also protect the connectors from corrosion or contamination by excluding metallic particles, moisture, and aircraft liquids. For information regarding the application of a sealing compound consult the current electronic material change on this subject.

Some electronic equipment is housed in a PRESSURE TYPE HOUSING. Many of these units use an air compressor to maintain a predetermined air pressure which may vary from sea level pressure to as much as 45 pounds per square inch absolute. The air in these components is dehydrated through the use of a desiccant (dehydrator). This pressurized dry air minimizes corona losses and prevents voltage breakdown when the equipment is operated at high altitude. When the equipment is used in moist climates or for frequent high altitude flights, the pressurization system will require more frequent inspection, such as checking the operation of the automatic pressure switches and the condition of the desiccant. (NOTE: The accurate and timely checking of the pressurization system becomes more important when it is realized that excessive pressure has caused units to explode at high altitude.)

Another means of preventing high-voltage arc-over when operating at high altitude is by automatically reducing the voltage in the critical circuits. This is accomplished by using a PRESSURE-SENSITIVE SWITCH.

Another maintenance consideration, in connection with the operation of equipment at extremely low temperature, is concerned with ELECTRIC HEATERS. Some equipments are provided with an automatic heater system which prevents mechanical overloading of rotating assemblies. This overloading is caused by the increased viscosity of their lubricants during frigid conditions. Typical units using such heaters are pressurization pumps and radar antenna assemblies.

Some units of test equipment are provided with an INTERNAL HEATER to prevent damage due to moisture. When the equipment is located in areas of high humidity it should be connected to a source of power at all times. Power is automatically applied to the heater when the equipment's power switch is turned off. If the equipment is not to be used for long periods, it should be stored in its transit case.

RADIO NOISE INTERFERENCE.—Suppression of radio interference is a task of first importance to all electronics personnel. The problem has increased in proportion to the complexity of both the electric system and the electronic equipment. The aircraft, the engine, the electric system, and the electronic equipment are involved in the problem. Almost every component of the aircraft is a possible source of radio interference, which is the main factor in preventing the operation of receivers at full sensitivity. All personnel concerned should be familiar with the problem of radio noise and how to eliminate it. The following information is needed:

1. What radio interference is.
2. Where the interference originates.

3. How it gets into equipment.
4. How to identify it.
5. How to suppress it at its source.
6. How to segregate its path of entry into a receiver.
7. How to prevent its entry into a receiver.
8. What considerations enter into the design of an interference-free equipment.
9. How to position and install electrical and electronic devices.

This information is presented in detail in *Elimination of Radio Interference in Aircraft,* T. O. No. 16–1–45. Some of this information will be presented briefly in the following pages.

SOURCES OF RADIO NOISE.—The origin or sources of radio noise are divided into three general groups.

ATMOSPHERIC STATIC, or "atmospherics," is a burst of RF energy caused by electrical discharges in the atmosphere. Although the frequency spectrum of atmospheric static is very wide, only frequencies in and below the high-frequency band propagate far enough to be very troublesome at long distances from the electrical disturbance. Therefore, UHF and VHF receivers are seldom troubled by atmospheric static. Reduction of such static is obtained by use of frequency modulation, directional antennas, and noise-limiter circuits. Frequency modulation is not used extensively in aircraft radio communication because of the bandwidth requirements.

PRECIPITATION STATIC is caused by the development of large static charges on the aircraft when it is flown through snow, rain, ice crystals, or dust clouds. An aircraft can build up a charge of several hundred thousand volts in a few seconds. The resulting high voltage gradients at extremities and sharp points exceed the breakdown strength of air and cause noisy corona discharges. The conventional radio antenna, which must stand away from the body of the aircraft to be of effective height, is exposed to high electric fields. This means that corona discharges occur first in the antenna system, the very place that is most sensitive to noise. Precipitation static is reduced by using a completely insulated antenna system—that is, by using highly insulated wire instead of bare wire, and by insulating all connections and supports for the antenna wire. Precipitation static is reduced also by eliminating all sharp metal projections from the aircraft and by installing DISCHARGERS, which quietly discharge accumulated static charges at a high rate. A discharger consists of silver-impregnated cotton wicking encased in a flexible plastic tube with an aluminum mounting lug. The many fine high-resistance fibers provide a multitude of discharge points. The resulting discharges are quiet up to very high currents. For detailed information on precipitation static refer to NavAer 16–1–518, *Handbook Installation and Maintenance Instructions, Anti-Precipitation Static System.*

MANMADE RADIO NOISE is caused by electrical transients which occur during the operation of electrical or electronic equipment. In brief, manmade radio noise will be generated whenever an electrical circuit is opened or closed abruptly, such as by a relay, commutator, or other make-and-break devices. A similar condition exists when large amounts of current are periodically and abruptly started and stopped, as in radar circuits. An electric spark is a generator of electrical disturbances which appear to cover the entire radiofrequency spectrum.

SUPPRESSION OF MANMADE RADIO NOISE.—Suppression of radio noise has advanced to the point where the proper application of available techniques will insure that receiving equipment installed in the aircraft will operate at optimum efficiency. The suppression or elimination of manmade radio noise is based on the premise that if manmade it can be man-corrected. Four types of suppression techniques are involved.

ISOLATION is the easiest and most practical method of radio noise suppression and revolves around the possibility of separating the source of radio noise from the input circuits of the receiving equipments affected. As every radio noise source can be considered a small transmitter, it is obvious that the radio noise source and leads carrying radio noise energy should be kept as far away from receiver antennas or lead-ins as possible. In many cases, the radio noise in a receiver may be entirely eliminated simply by moving the antenna lead-in wire just a few inches away from the source of radio noise. The value of sufficient separation between sources of radio noise and receiver input circuits is not apt to be overemphasized. The isolation method of radio noise suppression is very popular as it has the advantages of not requiring any additional material or adding any additional weight.

BONDING is a very necessary means of radio noise control. It provides grounding of all insulated conducting objects on the exterior of the aircraft. When conducting objects are not grounded, flight through precipitative weather conditions causes high-voltage charges to build up on those objects. Repeatedly, the voltage gets high enough to spark over to an adjacent ground member or the object discharges to the surrounding air by corona conduction. Either mode of discharge causes considerable radio noise.

Other important functions of bonding are to protect the aircraft and personnel from lightning discharges by equalization of potentials produced which might cause arcs and sparks in the aircraft structure, to provide a homogeneous counterpoise for radio transmission and reception, to provide power current return paths, and to provide a short path for bypassing RF noise. All electronic equipments should be grounded to the aircraft structure. This will be done by using short bond straps or by sheets of high conductivity (copper or aluminum) metal where it is impossible to use a short bond strap. No bond strap should be more than four inches in length.

SHIELDING is one of the most effective methods of suppressing radio noise. The primary object in shielding is to electrically "bottle up" the radiofrequency noise energy. In practical application, this means that the radio noise energy must be kept flowing along the inner surface of the shield. The use of good shielding is particularly effective in situations where filters cannot be used and are not particularly effective when they are used. A good example of this is where radio noise energy radiates from a radio noise source and the radiated energy is picked up by the various circuits that eventually connect to the receiver input circuits. It is obvious that it would be impractical to filter a number of leads or units that are influenced by the radio noise energy; hence, the application of effective shielding at the noise source itself is advisable for it will eliminate the radiated portion of the radio noise energy by confining it within the shield at its source.

Radio interference as radiated or conducted from a source may be of a single frequency or may cover an extended band of frequencies. When bonding, shielding, or isolation of the source proves ineffective as a means of reducing radio interference, it becomes necessary to employ filters to accomplish this reduction. A **FILTER** is defined as "a selective network which transmits freely electric waves having frequencies within one or more frequency bands and which attenuates substantially electric waves having other frequencies." The size of a filter may vary widely depending on the voltage and current requirements as well as the degree of attenuation desired. Filters are usually incorporated in equipment known to generate radio interference, but these filters are often inadequate, and in many cases it is necessary to add filters external to these equipments. This is especially true if the source of interference is coupling interference to paths of entry to a receiver other than the power line.

The types of filters used in the reduction of radio interference vary with the application, but each of the general filter types may be found to be particularly adaptable to some specific situation. Most of the electrical devices connected to power lines have features required for their operation, which are conducive to the generation of radio interference. The interference generated by these devices, unless properly attenuated, is impressed upon the power lines and conducted to the receivers. It may also be conducted into the receivers by inductive coupling to other wiring associated with the receivers. This interference, unless attenuated by means of filters, is then transmitted along these power lines, entering the receivers at the power line input; or this interference may be radiated somewhere along the power lines and enter the receiver by means of the antenna system.

Filters are of four kinds and are defined as follows:

LOW-PASS FILTER, which introduces negligible attenuation at all frequencies below a certain frequency, called the cutoff frequency, and relatively high attenuation at all higher frequencies.

HIGH-PASS FILTER, which introduces negligible attenuation at all frequencies above a certain frequency, called the cutoff frequency, and relatively high attenuation at all lower frequencies.

BAND-PASS FILTER, which introduces negligible attenuation at all frequencies within the range between two frequencies, and relatively high attenuation at all other frequencies.

BAND-ELIMINATION FILTER, which introduces negligible attenuation at all frequencies outside a certain range, and relatively high attenuation at all frequencies inside that range. (NOTE: For information that covers the theory of operation of these filters refer to NavPers 10087, *Basic Electronics*.)

The normal characteristics of a filter are obtained only when the filter is properly terminated in its characteristic impedance.

A WAVE TRAP is a filter or network especially designed to reject certain frequencies, or bands of frequencies. Networks of this type may be installed at the antenna of the transmitter or receiver in order to attenuate frequencies outside of the assigned frequency range of the equipment. All such networks must have low insertion loss, or attenuation, for the pass frequencies. In the design and construction of wave traps, the insertion loss is usually below 2 db.

There are two basic circuit configurations for filter networks, the pi-section and the T-section. Each may be broken down into half sections which have an inverted L-shape and are known as L-section filters. If a number of pi- or T-sections are connected in series to form a filter, the resultant network is called a LADDER NETWORK. Any of the above circuit configurations may be used for radio interference elimination.

In general, the use of simple CAPACITOR FILTERS is to be preferred over that of the more complicated network filters in cases where this type of filter provides the required degree of radio interference attenuation. In this method, the radio noise energy passes through the capacitor to ground and then back to its source. This short-circuiting effect is due to the fact that the capacitor offers a very low impedance path across the noise source terminals.

A given capacitor is effective in BYPASSING only a limited range of radio interference frequencies because of its internal inductance and the inductance of the connecting leads. The inductance of the capacitor depends upon its capacity, the material of which it is fabricated, and the length of the connecting leads. The capacitor leads are the major contributors to the inductance of capacitors. For these reasons, small mica capacitors with short leads are more effective as filters at high frequencies than large paper capacitors with normally long leads. ELECTROLYTIC CAPACITORS SHOULD NEVER BE USED AS FILTERS BECAUSE OF THE DANGER OF DIELECTRIC BREAKDOWN.

The popularity of the capacitor type filter is due to the fact that the current used for operation of the radio noise source does not have to pass through the filter. The only energy passing through the filter is the radio noise energy. The most important limiting factor in the choice of a capacitor-type filter is the breakdown voltage rating of the capacitor. It must be well above the voltage used to operate the source of radio noise to be filtered. For example, where a 24-volt source of noise is to be bypassed with a capacitor, the working voltage of the capacitor should be at least 50 volts.

TEST SETS FOR RADAR PERFORMANCE MEASUREMENTS

No matter for what purpose a radar is used, it should be operating at the efficiency for which it was designed. To insure that maximum efficiency is obtained, an overall performance check on the equipment is necessary.

Performance is influenced by the transmitter, because it determines the power and frequency characteristics of the output pulse. It is influenced by the condition and tuning of the transmission-line adjustments, which must be properly tuned, or attenuation through them may be great. It is also influenced by the tuning and the condition of the TR tube. In the preamplifier, mixer, and receiver, performance is influenced by the frequency and power of the local-oscillator input to the mixer, by the effectiveness of the RF to IF converter, and by the noise generated in the first RF tube or crystal and first IF tube. Subsequent receiver stages are not nearly so critical, because they amplify both signal and noise at relatively high levels and the noise they introduce is negligible compared with that already present. Other components of the radar system, (such as the modulator and synchronizer), must, of course, function properly but the minimum detectable echo pulse depends mainly on the quality of the receiver input circuit and transmitter power output. Overall performance will depend to a smaller extent on the PRF and the overall bandwidth of the receiver.

The relative influence on performance of the components mentioned above must be understood in testing a radar set. Ruinous losses are most likely to occur in the components that deal with the received signal from the time it is picked up by the antenna until it passes the first IF stage. Small troubles, bad connections, and lack of adjustment, unimportant though they may seem, may cause the complete loss of a signal as weak as that picked up from a small target. The receiving part of the radar is by far the most likely place for losses to occur that affect performance. Noise generated in the preamplifier and mixer cannot be reduced below some certain minimum. If a signal has dissipated too much of its energy before reaching the receiver, no amount of amplification will enable it to be seen above this noise on the indicator, since the noise is amplified with the signal. In the receiving system there are many opportunities for one-half or more of the signal power to be lost, and the pyramiding of these losses may make the performance extremely poor.

The frequencies used in radar cover a very wide range, from around 100 mc. to 30,000 mc. Therefore, a wide variety of test equipment must be used to perform the required maintenance in each designated frequency range.

Since the techniques used are different, dependent upon the range of frequencies involved, the technician should beware of the different types of test equipment that may be used. It is impractical to cover all of the test equipments in this text, therefore, only one typical test set, the TS–147A/UP, will be covered.

The use of an echo box to measure the overall radar system performance has been previously described in *AT 3 and 2,* NavPers 10317. A more detailed performance test may be made by checking each section of the radar system by means of a microwave signal generator, an oscilloscope, and suitable frequency and power-level meters. The TS–147A/UP combines most of these instru-

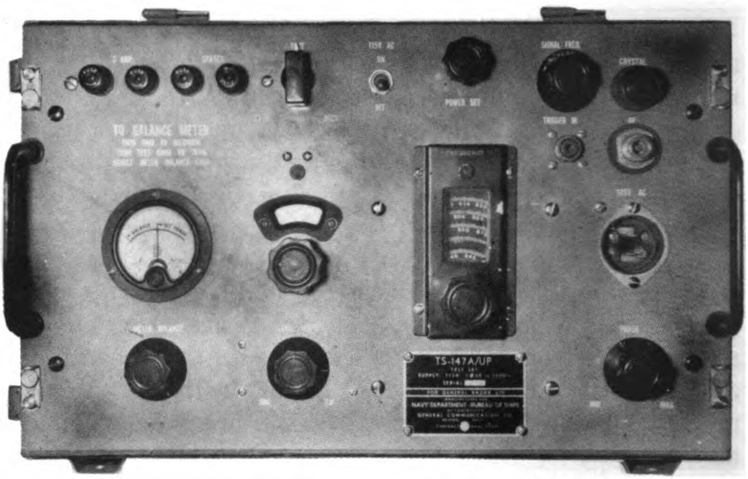


Figure 11–1.—Test Set TS–147A/UP.

ments in one unit and is designed for testing radar systems. This test set is shown in figure 11–1.

The TS–147A/UP Test Set

This is a portable microwave signal generator designed for testing and adjusting radar systems which operate within the frequency band of 8,500 mc. to 9,600 mc.

The test set supplies microwave test signals (either FM or CW) of known power level and frequency. These signals, together with the aid of an oscilloscope (A-type scan) for observing the radar video output, can be used in visual alinement methods in testing and adjusting radar systems. The sweep of the test signal frequency modulation and the horizontal sweep of the oscilloscope are triggered in synchronism with the radar system firing cycle. Because this timing makes the horizontal A-scope sweep proportional to the frequency of the FM test signals, the resulting patterns seen on the A-scope represent the frequency response curve of the receiver section of the radar system. The oscilloscope patterns show the effect of system adjustments.

The signal generator consists essentially of a reflex klystron oscillator and a triggered sweep generator which can modulate the frequency of the klystron by applying a sawtooth voltage to the reflector of the tube. The sweep can be triggered either by RF pulses applied to the RF receptacle or by voltage pulses applied to the TRIGGER receptacle of the test set.

Frequency modulated pulses are produced by an internal sawtooth sweep which can be varied to produce frequency excursions from zero (CW) to at least 60 mc. by changing the sweep amplitude. At maximum amplitude the sweep voltage covers three modes of the klystron. On the A-scope each mode covered appears as a sharp pip. The pips can be widened into broad response curves by decreasing the sweep amplitude. Changing the d-c level of the sweep changes the position of the test signals along the range axis of the A-scope. Test signals can be seen on all types of radar indicators as artificial signals.

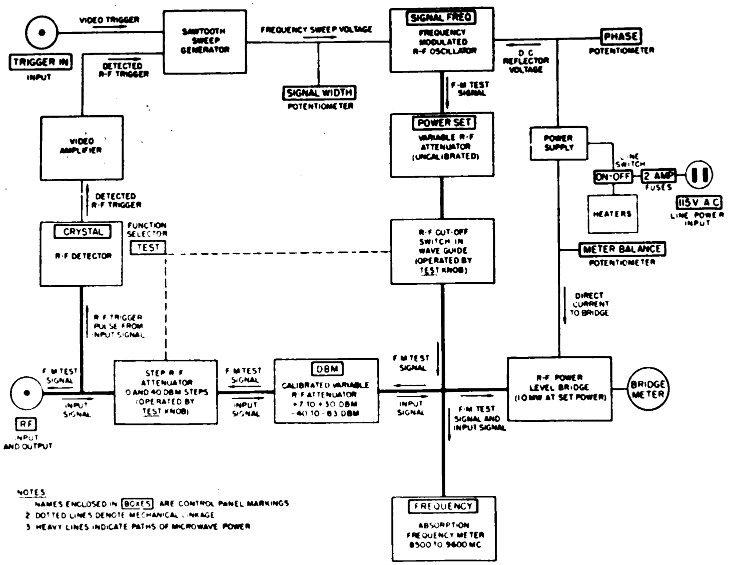
Average power measurements are made by a temperature compensated thermistor bridge wattmeter and a calibrated RF attenuator. The test set can measure both the power level of the test signals generated internally and the power level of external signals applied to the test set. Power is measured in dbm (decibels relative to one milliwatt) over a range of minus 42 to minus 83 dbm for peak levels of test signals supplied by the set and plus 7 to plus 30 for average levels of external power applied to the test set. These measurements are accurate within 2.0 db. The ranges stated give the levels at the RF input-output connection on the test set, and in use the attenuation of cables and coupling devices must be added to these levels.

Frequency measurements are made by an absorption frequency meter which uses the thermistor bridge wattmeter as a tuning indicator. When tuned to resonance, the frequency meter reflects part of the power so that a dip in the wattmeter indication is observed. The frequency meter can be used both for test signals generated in the test set and external power applied to the test set.

THEORY OF OPERATION.—A block diagram of the test set components and their interconnections is shown in figure 11–2. (NOTE: For ease of understanding the control panel, markings are enclosed in boxes in the block diagram and they are capitalized in the text.)

The test signals are generated by the test oscillator which is frequency modulated by the sawtooth voltage supplied by the triggered sweep generator. The sweep generator is triggered in synchronism with the system firing cycle either by means of voltage pulses connected to the TRIGGER IN (input) receptacle or by means of pulse RF power detected and amplified by the detector crystal and trigger amplifier.

The output of the test oscillator is controlled directly by means of the POWER SET attenuator and the RF cutoff. The POWER SET attenuator provides adjustment of the signal level applied to the thermistor bridge. The RF cutoff is a waveguide switch which serves as a means for turning the output of the oscillator on or off.



NOTES
1. NAMES ENCLOSED IN BOXES ARE CONTROL PANEL MARKINGS
2. DOTTED LINES DENOTE MECHANICAL LINKAGE
3. HEAVY LINES INDICATE PATHS OF MICROWAVE POWER

Figure 11–2.—Block diagram, TS–147A/UP.

The output level of the test signal is measured by the DBM attenuator. When the POWER SET attenuator is adjusted so the thermistor bridge is balanced, the DBM dial indicates directly the test signal power level at the RF receptacle. The step attenuator, operated by the TEST (function selector) knob, sets the range of the DBM attenuator.

The frequency of the test signal is measured by the frequency meter. When the frequency meter cavity is tuned to the test signal frequency, part of the power from the test oscillator is reflected and a dip in the bridge meter reading is produced. The reflection of the frequency meter also causes a decrease in the power output of the test set. If the test signal is frequency modulated and the frequency meter is tuned within the range of the frequency sweep, power will be absorbed each time that the frequency sweep reaches the resonant frequency of the frequency meter. The instantaneous decrease in power output resulting from the absorption will appear as a small negative pip in the frequency response curve seen on the A-scope.

In addition to operating as a signal generator, the test set also measures the power and frequency of external RF signals applied to the test set. Power is measured by means of the DBM attenuator and the thermistor bridge wattmeter. For this measurement the TEST knob is set to TRAN position to close the RF cutoff and set the DBM attenuator to the proper range. The DBM attenuator is then adjusted to decrease the input power to the level which balances the thermistor bridge wattmeter. At this point, the power level at the bridge is 1.0 milliwatt and the DBM dial indicates directly the input power level in dbm (decibels above 1.0 milliwatt).

The frequency of input signals is measured by the same means used for test signals. At resonance the frequency meter absorbs part of the power applied to the thermistor bridge and causes a dip in the bridge meter reading.

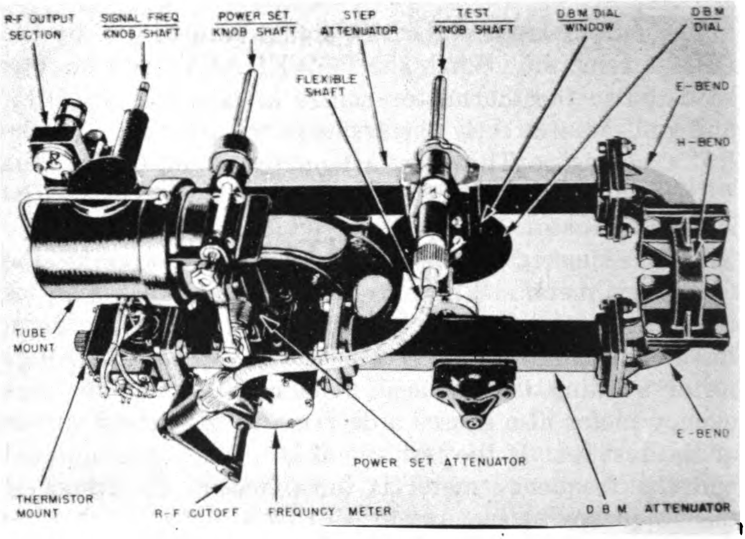


Figure 11–3.—RF plumbing arrangement.

Figure 11–3 shows the physical layout of the RF PLUMBING COMPONENTS in the test set. The plumbing components are waveguide sections fitted with attachments or modifications to vary the way in which RF power is transmitted through them. Most of the components are fixed assemblies which cannot be disassembled in the field for maintenance or repair. Some components are calibrated assemblies which, if taken apart in the field, can no longer effect accurate measurements. In any case, however, each component is replaceable as a complete unit which can be easily removed or inserted in the set.

The arrangement of the plumbing comprises three branches; namely, the input (or output) branch, the oscillator branch, and the bridge branch. The three branches join in a T-section at which junction the wavemeter is located.

The TRIGGER DETECTOR AND AMPLIFIER CIRCUIT (fig. 11–4) provides the means for triggering the frequency sweep generator from pulse RF power applied to the RF

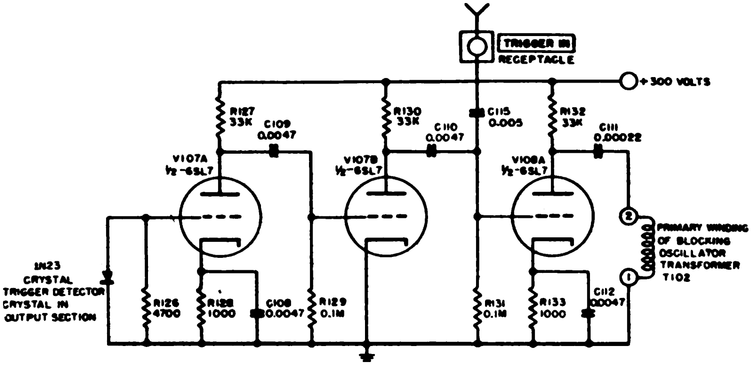


Figure 11–4.—Trigger detector and amplifier circuit.

receptacle. The RF pulses are detected by the crystal in the output coupling section and the resulting pulse voltage is applied to the input of the trigger amplifier. The amplifier is a three-stage resistance-coupled amplifier. The voltage gain is approximately 500. The output of the amplifier supplies a voltage pulse which can trigger the blocking oscillator sweep circuit. The output pulse is parallel fed through *C*111 to the blocking oscillator transformer, *T*102.

Video triggers applied to the TRIGGER receptacle of the test set are amplified by the last stage of the amplifier. The voltage trigger pulses feed through *C*115 directly to the grid of *V*108*A*.

The FREQUENCY SWEEP CIRCUIT for the test oscillator is a sawtooth generator of the triggered blocking-oscillator type. (See fig. 11–5.) The blocking oscillator consists of *V*108*B* and a three-winding transformer, *T*102. The circuit comprises a conventional blocking oscillator except that the grid is biased below cutoff so that a trigger pulse must be applied to start each cycle oscillation. The grid bias is applied to the cathode by the voltage divider *R*134 and *R*135.

The blocking oscillator generates sawtooth pulses by discharging *C*114. When a trigger pulse is applied to the transformer *T*102, the grid is raised above cutoff and

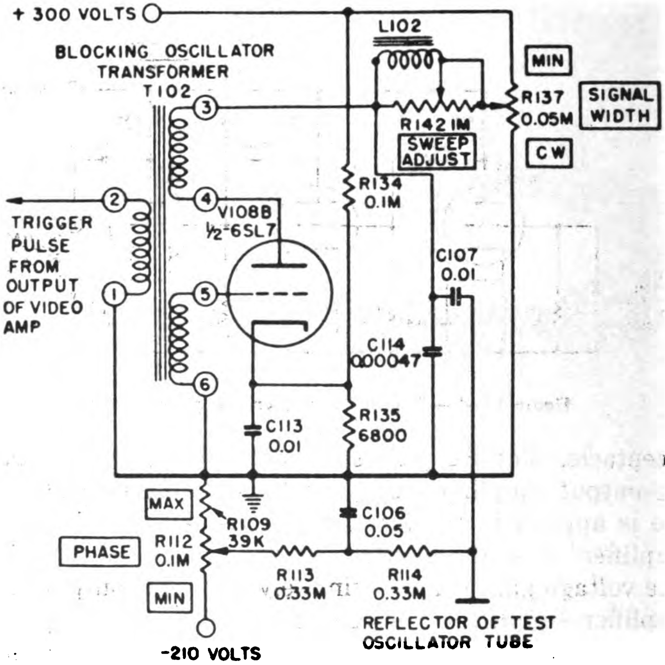


Figure 11–5.—Blocking-oscillator sweep circuit.

oscillation starts. The plate current which flows during the time the grid is above cutoff causes a voltage drop across the plate resistor (*R*142 and part of *R*137). This decreases the voltage across *C*114 to a value below the steady state voltage. When the single cycle oscillation ceases, the plate current is zero but a voltage drop still exists in the load resistor *R*142 due to the charging current of *C*114. This voltage decreases to zero as the capacitor *C*114 charges to the steady state voltage. The charging curve which occurs after the oscillation ceases is the sawtooth sweep used to modulate the test oscillator.

The amplitude of the sweep voltage depends upon the steady state voltage across the capacitor *C*114. The voltage is made adjustable from 0 to plus 300 volts by means of the SIGNAL WIDTH potentiometer *R*137. As the steady state voltage is increased by the SIGNAL WIDTH control, the amplitude of the sweep voltage is increased. The limits of the control are marked CW and MIN on the control panel. CW corresponds to zero voltage and MIN corresponds to maximum voltage. These reference marks designate the effect of the SIGNAL WIDTH control on the oscilloscope pattern obtained during receiver tests.

The TEST OSCILLATOR TUBE used in the test set is a type 2K25 reflex klystron. There are two methods for tuning a reflex klystron tube. It can be tuned mechanically by varying the volume of the resonator cavity and spacing of the resonator grids. The tuning range is covered by approximately three and one-half turns of the SIGNAL FREQ knob and allows a variation of approximately plus or minus six percent about a specified frequency. The second method of tuning the oscillator is by varying the reflector voltage.

As reflector voltage is varied, the tube will oscillate in different modes. The relation of power output and frequency to reflector voltage is shown for three modes in figure 11–6. The curves indicate that the power output of the tube is greater in the modes with more negative reflector voltages. The range of frequency sweep within each mode decreases with increasingly negative reflector voltages. If the sawtooth voltage is applied to the reflector about the center value of 110 volts (d-c level), the frequency of the oscillator will be swept through one or more modes and the power output will be frequency-modulated. The number of modes covered depends upon the amplitude of the sawtooth. This is determined by the setting of the SIGNAL WIDTH control.

When the d-c level of the sawtooth is changed to a lower voltage, the sawtooth will sweep over a lower mode. Also the time at which the center frequency of the mode is reached occurs later. Such a change appears as a shift in the position of the frequency response curve along the range axis of the A-scope. The d-c level of the sawtooth is controlled by the PHASE control.

The THERMISTOR BRIDGE CIRCUIT used in this test set is

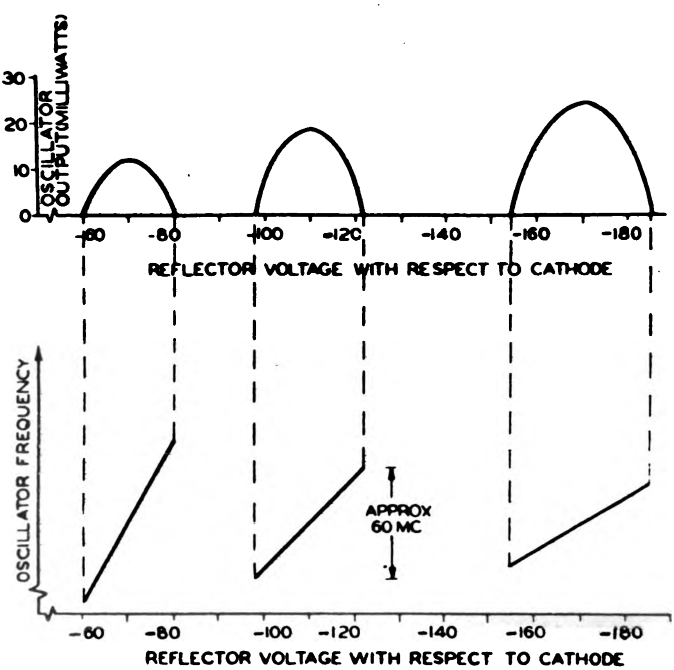
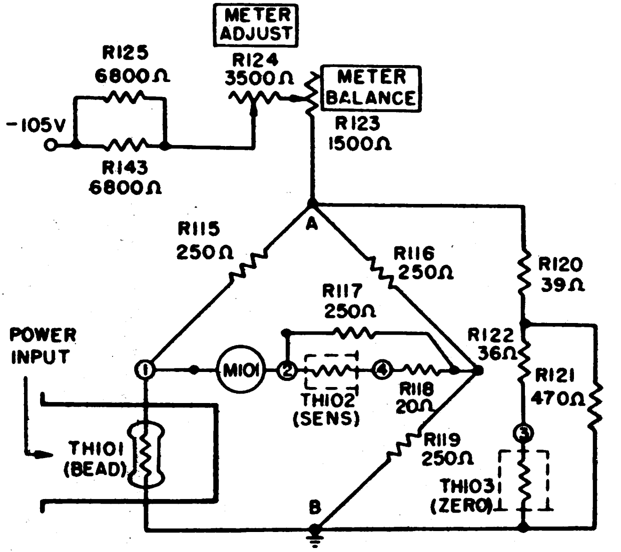


Figure 11–6.—Reflex klystron characteristics.

covered in detail since this circuit finds frequent use in other test equipments. The thermistor bridge wattmeter circuit shown in figure 11–7 is a Wheatstone bridge which measures the resistance of an RF power sensitive resistor called a thermistor. The thermistor has a negative temperature coefficient of resistance; i.e., as the temperature is increased the resistance decreases. The temperature, and therefore the resistance, of the thermistor is determined by the temperature of its environment, by the current flowing through it, and by absorbed RF power. If the current and ambient temperature conditions are properly controlled, RF power can be measured by measuring the change in thermistor resistance caused by the applied power.

There are two types of thermistors used in the test set. The BEAD THERMISTOR (fig. 11–8 (A)), used for



TH101= BEAD THERMISTOR
TH102= DISK THERMISTOR (SENSITIVITY COMPENSATION)
TH103= DISK THERMISTOR (ZERO COMPENSATION)

Figure 11–7.—Complete thermistor bridge circuit.

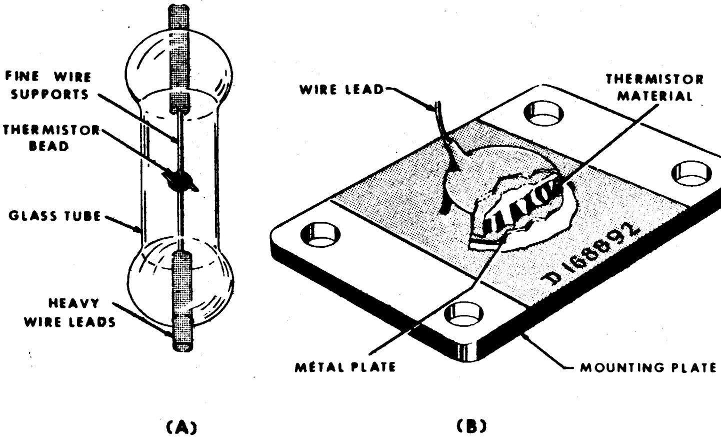


Figure 11–8.—Thermistors.

power measurement, has a small mass and is affected by the changes just mentioned. The DISK THERMISTOR (fig. 11–8 (B)), used in compensating networks, has a large mass. Its resistance is relatively insensitive to the current flowing through it and is dependent primarily on the ambient temperature. A simplified thermistor bridge circuit is shown in figure 11–9. The Wheatstone bridge consists of three resistors, $R115$, $R116$, and $R119$, and the bead thermistor. When the Wheatstone bridge is balanced, no current flows in the meter. The bridge is balanced when the resistance meets the following condition:

$$\text{Resistance of thermistor} = R115 \times \frac{R119}{R116}$$

In the test set, the bridge is designed so that it will balance when one milliwatt of RF power is applied to the bead thermistor. This is done by adjusting the initial thermistor resistance so that, when one milliwatt of power is applied, the resistance decreases just enough to balance the bridge.

The initial setting of the thermistor resistance is made by adjusting the current through it. This is accomplished by the METER BALANCE potentiometer, $R123$. The constants of the bridge are chosen so that the required initial value of thermistor resistance produces 150 μa. through the meter. This current is marked BALANCE on the meter scale (on some test sets it is called ZERO SET).

The initial adjustment of current is referred to as "balancing" the meter because, when the adjustment is made, the meter reads BALANCE. It must be realized, however, that in this condition the bridge is UNBALANCED by a known amount. Balance of the bridge is not attained until the meter current is reduced to zero by applying one milliwatt of power to the bead thermistor. The zero current point on the meter scale is marked SET POWER.

This basic thermistor bridge circuit (fig. 11–9) must be modified in practical operation to overcome two dis-

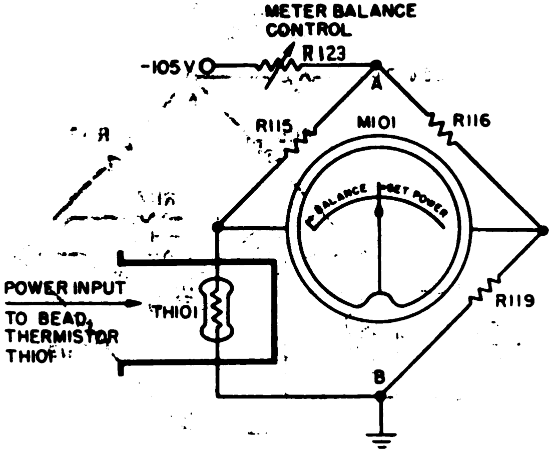


Figure 11–9.—Basic thermistor bridge circuit.

advantages. Means are employed to compensate for the fact that the meter sensitivity (microamperes of meter current per milliwatt of RF power) varies with temperature and that because of changes in the bead thermistor ambient temperature, the METER BALANCE control must be readjusted before every measurement. The methods employed to compensate for these disadvantages will now be explained.

The thermistor bridge (fig. 11–9) increases in sensitivity with a decrease in temperature. The sensitivity (ohms watt) of the thermistor bead remains constant over a broad temperature range, but the bridge sensitivity (microamperes meter reading per milliwatts RF power) increases with a decrease in temperature. As the temperature goes down, the thermistor resistance tends to go up, making it necessary to apply a higher voltage to the bridge by adjusting the METER BALANCE control, *R*123. Because of this higher voltage, the change in the current through the meter, due to a given change in thermistor resistance, is greater at lower temperatures.

This variation in sensitivity can be compensated for by increasing the meter resistance as the temperature de-

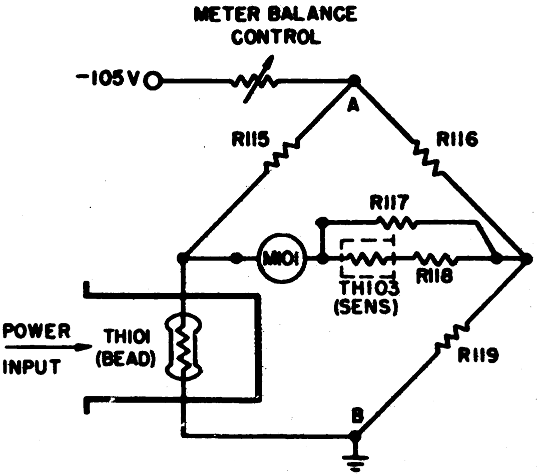


Figure 11–10.—Basic thermistor bridge circuit with sensitivity compensation.

creases. Since thermistors have this characteristic of increase in resistance with decrease in temperature, a disk thermistor is placed in series with the meter. (See fig. 11–10.) It is necessary to put resistor *R*118 in series and resistor *R*117 in parallel with the disk thermistor to obtain the correct rate of change of resistance with temperature so that the sensitivity of the bridge remains constant over the required temperature range.

To eliminate frequent resetting of the METER BALANCE control, an automatic method is provided for increasing the current through the bead thermistor with a decrease in temperature. A disk thermistor whose resistance increases with decreasing temperature is placed in parallel with the bridge. (See fig. 11–11.) Since more current flows through the bridge at lower temperature, the ratio of the resistance of the shunting disk thermistor to the resistance of the bridge (across points *A* and *B*) is greater at lower temperature. To adjust this compensation to cover the required temperature range, the disk thermistor is combined in a network with resistors *R*120, *R*121, and *R*122.

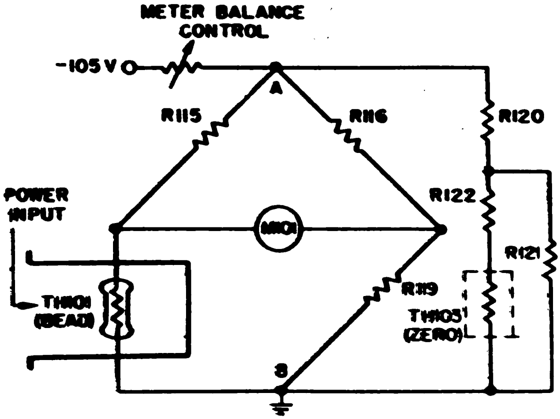


Figure 11–11.—Basic thermistor bridge circuit with zero compensation.

The thermistor bridge used in the test set has both sensitivity compensation and zero drift compensation. Resistance *R*124 (METER ADJUST) is added in series with the METER BALANCE potentiometer *R*123 to adjust the range of the METER BALANCE control to suit the particular thermistor that is used.

The compensating networks are designed for an average bead thermistor so that the compensation is correct at 0° C., 30° C., and 60° C. At intermediate points, the compensation is not exact although it is satisfactory for practical use. These networks are not perfect for any bead thermistor due to the variation in electrical characteristics of bead thermistors. As a result, the manual METER BALANCE control cannot be eliminated entirely.

OPERATION AND ADJUSTMENT.—This test set is a portable instrument which requires no installation or special adjustments before it is connected to a power supply and put into operation. When making connections the operator should arrange the equipment so that it is possible to operate the test set controls and observe the patterns on the oscilloscope screen while making adjustments on the radar system.

The test set will function from a power source of 50 to 1,600 c.p.s., thus is may be operated from a shop or aircraft power source. The performanee test as given in the instruction book is for the purpose of ascertaining if the set is operating properly.

The function of the operating controls, in summary form, follows:

1. Power switch applies power to the gear and turns the internal heater off.
2. METER BALANCE adjusts the current through the meter.
3. DBM measures the average level of transmitted power in decibels above one milliwatt and receiver sensitivity below one milliwatt.
4. TEST (FUNCTION SELECTOR)
   a. TRANS position sets the range of DBM for transmitter power measurements and cuts off the entire receiver section.
   b. TEST (DOT) position turns on the test signal and sets the range of the DBM for high level signal outputs.
   c. RECV position turns on the test signal and sets the range of the DBM for low level signal outputs.
5. FREQUENCY tunes the absorption type wavemeter which is the resonant cavity.
6. POWER SET adjusts initial power level of the test signal to balance the bridge.
7. SIGNAL FREQ tunes the klystron oscillator to the desired frequency.
8. SIGNAL WIDTH sets the amplitude of the sawtooth sweep applied to the klystron and therefore the amount of FM from zero (CW) to over 40 mcs.
9. PHASE controls the d-c reflector voltage of the klystron.

The use of this test set in the full scope of its applications requires four CONNECTIONS:

1. Power must be supplied to operate the test set.
2. RF coupling must be made between the test set and the radar system. Either the pickup horn or the directional coupler can be used to couple the cable to the radar system.
3. Trigger for the test set sweep generator is required when the FM test signals are to be used. A trigger is supplied by the RF pulses applied to the RF receptacle without any additional connections or by any standard video trigger, such as IFF trigger, applied to the TRIGGER receptacle of the test set.
4. An oscilloscope (AN/USM–24) connected to the video output of the radar receiver is used to indicate the response of the system by A-scan.

(NOTE: The last two connections are not required for checking a radar transmitter.)

MAINTENANCE.—Because the test set is a calibrated measuring instrument, field maintenance is necessarily limited to troubleshooting in the electrical circuits and to replacement of complete assemblies in the plumbing. At all times, field maintenance personnel must take care to avoid disturbing the fixed adjustments on any calibrated assembly. If calibration of any part is changed, then the only practical field procedure is to replace the entire assembly with a factory calibrated unit. General inspection and test of the set can be performed by following the check described in the *Handbook of Service Instructions.*

The test set contains only two internal adjustments which can be made by operating personnel without danger of changing its calibration. These are the meter adjust potentiometer and the sweep adjust potentiometer.

MEASURING RADAR PERFORMANCE.—The primary purpose of this test set is to measure the performance of a radar system. The minimum performance standards will vary from one type of radar to another. The minimum standards for a radar may be obtained from the *Hand-*

book of Service Instructions for the particular equipment. The accuracy of performance measurements is important. For example, in a system whose performance is down 5 db its maximum range will be only 75 percent of rated performance.

A list of HELPFUL HINTS ON RADAR TROUBLESHOOTING follows:

1. It is generally advisable to take all measurements on all ranges of operation of a particular radar under test.
2. Always adjust the radar's crystal current before attempting a sensitivity check.
3. The frequency of magnetrons usually increases with a decrease in pulse width.
4. Recovery time decreases with a decrease of pulse width.
5. Recovery time decreases with an increase in receiver sensitivity.
6. Receiver sensitivity increases with a decrease in bandwidth.
7. Receiver sensitivity usually decreases with a decrease in crystal current.
8. Magnetron frequency decreases with an increase in reflected load from the antenna.
9. Frequency pulling by the antenna must be checked with the antenna in motion.
10. A magnetron which arcs under test can be the result of:
    a. The magnetron not conducting hard enough.
    b. Not enough warmup time.
    c. Low filament voltage.
    d. PRF too low.
    e. Pulse width too narrow.
    f. Excessively high or low line voltage.
11. Normally, returning the magnetron to standby to

allow longer warmup time will stop the magnetron from arcing.

12. Local oscillators (klystrons) in radar receivers are usually operated above the magnetron in search and below the beacon station on beacon.
13. Intermittent troubles can usually be located effectively by placing several pieces of test equipment at several points in the radar under test. For example, intermittent targets. Possible causes could be crystal, duplexer, magnetron, IF strip, detector, video amplifiers, indicator, power supplies, or the antenna. In this case where any one could very easily be the cause of the trouble, the following checks will help isolate the trouble:
    a. A test scope on the video detector will show whether the trouble is the result of the magnetron, klystron, IF strip, detector, antenna, or crystal. If the output here is constant, the trouble is in the video amplifiers or the indicator.
    b. If the output at the detector is not constant, a meter connected to read magnetron current will show whether the magnetron is at fault. While watching the scope, see if the magnetron current is fluctuating as the targets are lost.
    c. If the antenna is suspected, the targets will usually be intermittent at the same azimuth setting. If it is found that the targets do become intermittent at the same azimuth position of the antenna, the rotating joint will usually be the cause.
14. Low output power may be the result of:
    a. Low line voltage.
    b. Bad high voltage rectifiers.
    c. Poor magnetron.
    d. Pulse width too narrow.
    e. PRF too low.
    f. Weak magnetron.

15. Erratic AFC operation other than the AFC units themselves may be the result of:
    a. The antenna pulling the magnetron frequency, causing the AFC to fall out.
    b. Magnetron frequency modulation.
    c. Erratic operation of the local oscillator.
    d. The AFC being set up on the slope of the klystron curve instead of at the top.
16. A little common sense, a basic knowledge, and test equipment makes troubleshooting a radar as easy as troubleshooting a radio.

While power is important in determining maximum range, system gain is much more so, since the effect of the gain in increasing range is equal to the SQUARE of the effect of the power in increasing the range. It is therefore important for the technician to keep the gain, or sensitivity, of the radar at the peak of possible performance. The radar receiver should be centered at the transmitter center frequency, and the local oscillator must be made to operate at the magnetron frequency plus (or minus) the IF center frequency. The AFC circuits must be accurately alined to prevent loss of valuable system gain.

TEST SETS FOR PULSE EQUIPMENT MAINTENANCE

Many radar timing circuits are designed to produce electrical pulses of precise duration and of distinctive waveform. It is therefore obvious that a device by which waveforms can be observed and measured is a valuable instrument in corrective maintenance. These distinctive waveforms are provided in most *Handbook of Service Instructions*. The technician should check all test points for which waveforms are given. He should check then when the equipment is operating perfectly and should note any discrepancies between the operating equipment and the instruction book.

A typical oscilloscope that is capable of performing the just mentioned measurements is the AN/USM-24. An oscilloscope of this type is frequently referred to as a SYNCHROSCOPE since it is capable of producing a sweep of very short duration which is generated only when a synchronizing signal is provided.

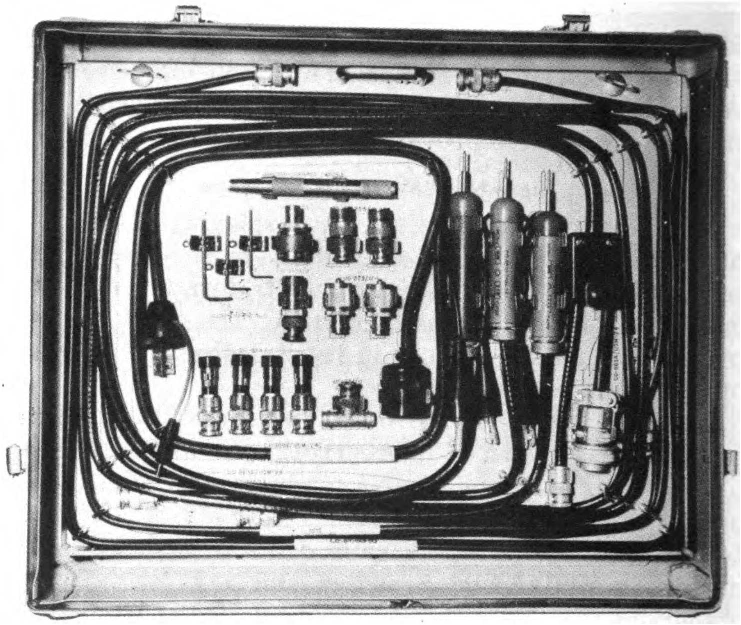
The AN/USM–24 Oscilloscope

This is a portable test set for displaying a luminous plot of the time variation of a voltage pulse or wave with self-contained means for measuring its duration, displacements, and instantaneous magnitude. The oscilloscope is primarily intended for use in bench testing of all types of electronic equipment.

The Oscilloscope AN USM–24, which is the complete equipment, is shown in figure 11–12 (A) and (B).

Since it is usually desirable to observe the instantaneous magnitude of electrical signals against time, the horizontal presentation is normally a linear time base. By means of a SWEEP μS switch and a FINE SWEEP potentiometer, the time base may be varied at will within the limits of 1.25 to 125,000 μsec. Also any portion of the time base over 10 μsec. may be selected and expanded for more detailed observation of the signal. The linear time base can be operated either in a trigger or a periodic fashion. The H GAIN control is provided to vary the horizontal length of the trace so that direct interpolation between markers and graph screen permit observation of time periods less than those provided for by markers.

Signal amplification is uniform over a wide frequency band so that flat-topped pulses may be adequately displayed. The signal is applied as vertical presentation. The rise time of the vertical channel is 0.04 μsec., and the tilt of a 5,000-μsec. flattop wave is less than 5 percent. Sensitivity of the vertical amplifier is such that signals with peak values from 0.01 to 150 volts may be observed directly. Two types of test probes, as accessories, are supplied. One enables signals up to 600 volts to be observed; the other enables signals up to 2 volts to be observed with only 9-$\mu\mu$f. shunt capacity loading. The ver-



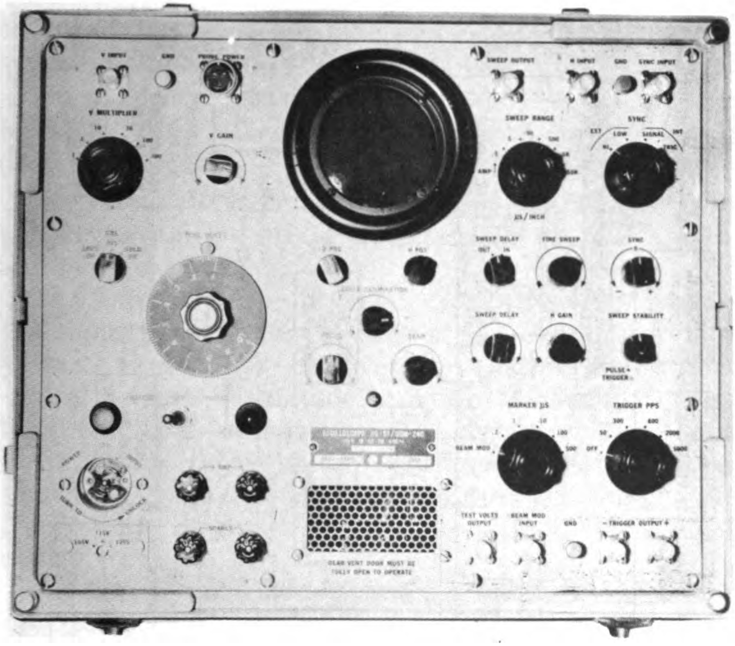
(A) OSCILLOSCOPE COVER CW268/USM-24

Figure 11–12.—Oscilloscope AN/USM–24.

tical channel includes a delay line of 0.055 μsec., thus making it possible to observe the leading edge of a pulse triggering the linear time base.

For observation of high-speed transients, the electron beam is only energized during "go" time of the linear time base, thus providing highlight brightness of the trace, without injury to the cathode-ray tube screen. The linear time base produces a gate which intensifies the beam during "go" time only. Also, time markers generated internally from 0.2 to 500 μsec. produce further intensification for accurate time measurement.

The oscilloscope has certain important auxiliary features, a calibrating generator for accurate measurement of instantaneous values of signal without recourse to external standards, a choice of five time markers from 0.2 to 500 μsec. for accurate time measurements, and a trig-



(B) OSCILLOSCOPE OS-26/USM-24

Figure 11–12.—Oscilloscope AN/USM–24. Continued.

ger generator with five fixed ranges from 50 to 5,000 pulses per second for triggering both the oscilloscope and external equipment.

All operating controls and connections to the oscilloscope are either on the front panel or accessible through the ventilating door on the back of the case. All accessories are mounted on the tray of the oscilloscope cover. Controls and connectors are arranged functionally and are so labeled that the oscilloscope may be operated without reference to the instruction manual.

The oscilloscope is designed to operate between minus 54° C. (minus 65° F.) to plus 65° C. (plus 150° F.) under humidity conditions up to 95 percent. However, under humidity conditions in the range of 75 to 95 percent the heater should be on when the oscilloscope power is not turned on. The oscilloscope may be used in altitudes up

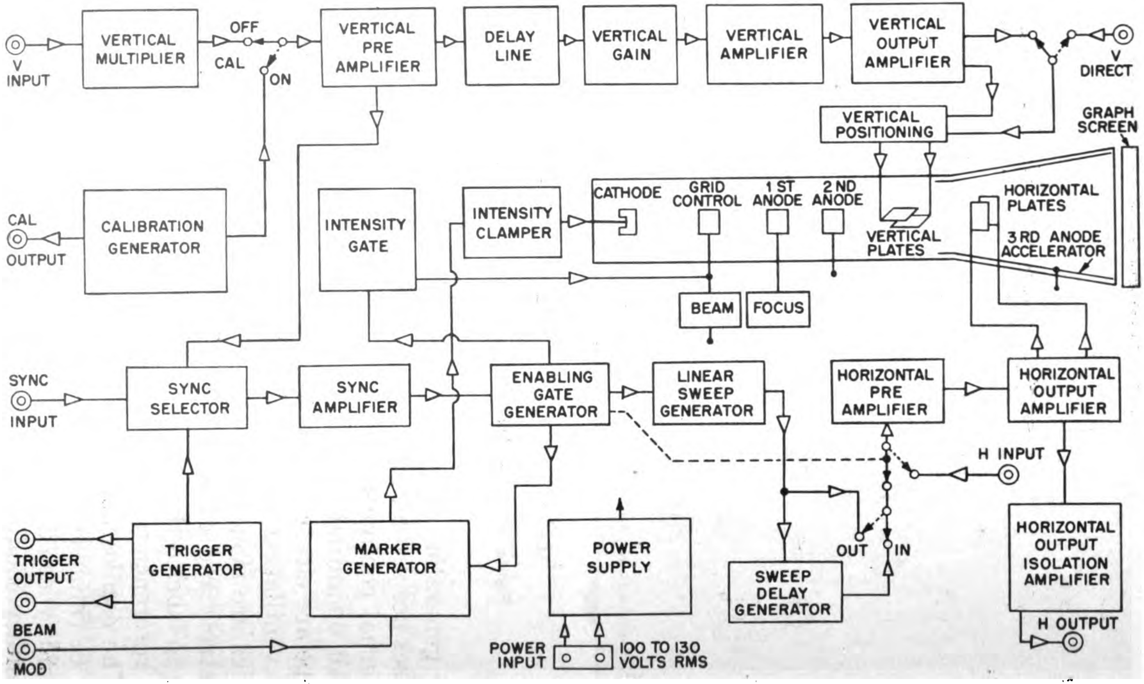


Figure 11-13.—Oscilloscope AN/USM-24, block diagram.

to 10,000 feet, and will take any reasonable shock and vibration imposed on it by normal usage.

PRINCIPLES OF OPERATION.—The oscilloscope consists basically of the following elements, each of which will be described briefly.

The DISPLAY CHANNEL consists of the cathode-ray tube and its associated controls which regulate focus, intensity, and picture centering.

The function of the VERTICAL CHANNEL is to transmit the signal from the input back to the cathode-ray tube with no appreciable change in its wave form but with control over its magnitude in order to handle properly the wide range of signals to be observed. The signal must also be delayed by sufficient time, so that the linear time base, markers, and intensification gate may start functioning properly before the signal reaches the cathode-ray tube.

As shown on the block diagram (fig. 11–13), the vertical channel includes several stages of amplification, means for varying the gain of the channel, a calibrated multiplier for attenuating the signal, and a network for delaying the signal. Signals from 0.10 to 150 volts can be coupled directly to the input jack. Signals may be connected to the deflection plate of the cathode-ray tube through the V DIRECT terminal on the back of the oscilloscope. The RC network of the V MULTIPLIER network is such that the input impedance is constant and is equal to 300,000 ohms resistance shunted by a 37-$\mu\mu$f. capacitance. The voltage divider is frequency compensated so that no distortion of input signal results within the pass band of the vertical channel. The V MULTIPLIER switch has six positions—times 1, 3, 10, 30, 100, and 300. Voltage as read by means of the CAL potentiometer is to be multiplied by the setting of the V MULTIPLIER switch. When the 600-volt probe is used, the reading should be multiplied by ten.

The input signals are amplified in the VERTICAL PREAMPLIFIER where they are coupled to the SYNC

SELECTOR and also to the VERTICAL AMPLIFIER through a delay line. The remainder of the VERTICAL CHANNEL consists of the VERTICAL AMPLIFIER and the VERTICAL OUTPUT AMPLIFIER. These are conventional circuits found in all oscilloscopes.

The HORIZONTAL CHANNEL consists of the HORIZONTAL PREAMPLIFIER, HORIZONTAL OUTPUT AMPLIFIER, and HORIZONTAL ISOLATION AMPLIFIER. The input to the HORIZONTAL CHANNEL is either from the TIME BASED CHANNEL or the external H INPUT jack. The HORIZONTAL PREAMPLIFIER and the HORIZONTAL OUTPUT AMPLIFIER are conventional. The HORIZONTAL ISOLATION AMPLIFIER prevents loading at the H OUTPUT jack from affecting oscilloscope operation.

The INTENSITY CHANNEL provides the modulation of the electron beam. The intensity gate circuit receives a positive gate from the ENABLING GATE GENERATOR. This circuit provides a positive-going gate which is applied to the grid of the cathode-ray tube and a negative marker which is applied to the cathode. The constants of the circuit are such that the tube acts as a limiter, thus feeding a constant amplitude gate to the grid of the CR tube.

The TIME BASE CHANNEL consists of an ENABLING GATE GENERATOR which produces a gate whose duration may be varied from 1.2 to 120,000 μsec., a LINEAR SWEEP GENERATOR which produces a linear sweep as controlled by the ENABLING GATE GENERATOR, and a SWEEP DELAY GENERATOR which can be used to take any 10 percent portion of the linear sweep and magnify it to fill the screen of the cathode-ray tube. The ENABLING GATE GENERATOR can be operated either in trigger or repetitive condition, depending upon the setting of the SWEEP STABILITY control.

The SYNCHRONIZATION CHANNEL provides means of synchronizing the linear time base with the signal so that the trace appears stationary on the face of the cathode-ray tube. The channel consists essentially of a selector switch for selecting either external sync from the SYNC INPUT jack, internal trigger from the TRIGGER GENERATOR, or a sync signal as developed in the VERTICAL PREAMPLIFIER, thus the linear time base can be synchronized with either an external triggering pulse or with the internal trigger generator or with the internal signal.

The MARKER CHANNEL provides markers between 0.2 and 500 μsec. for the purpose of calibrating the linear time base. The MARKER GENERATOR consists basically of a cathode driven multivibrator, a gating tube, a marker amplifier, and a means of selecting markers. The multivibrator is driven by negative pulses from the ENABLING GATE GENERATOR. The marker amplifier is also used when external marker signals are applied through the BEAM MOD jack.

The TRIGGER CHANNEL is basically a pulse generator whose occurrence rate may be varied from 50 to 5,000 pulses per second. It consists essentially of a free running multivibrator that determines repetition rate and a blocking oscillator that produces trigger pulses. The purpose of the trigger generator is to provide a trigger for the oscilloscope and external equipment producing a waveform to be studied. The pulse output is 1.2 μsec. in duration, with a rise time of 0.1 μsec. and an amplitude of 45 volts. Two TRIGGER OUTPUT jacks are provided (one positive and one negative) ; a portion of the positive pulse is fed to the sync selector switch.

The CALIBRATION CHANNEL consists essentially of the CALIBRATION GENERATOR which is a 5-kc. symmetrical multivibrator. Its function is to provide a standard signal to the vertical amplifier for calibrating the incoming signal. Its additional function is to provide a signal of high level for checking and recalibrating, if necessary, the vertical attenuator as well as the auxiliary probes. (NOTE: Since the CAL VOLTS dial is calibrated in peak-to-peak volts, it is necessary to divide by 2.83 to convert to r.m.s. voltages.)

The POWER SUPPLY CHANNEL provides all the necessary voltages required for the operation of the oscilloscope as well as power for the ventilating system.

Generally speaking, the maintenance of this test set follows those procedures given in connection with Test Set TS–147A/UP (which has already been given in this chapter), and they will not be repeated. For detailed maintenance instructions consult the instruction book that is located in the oscilloscope cover.

The TS–375/U Vacuum-Tube Voltmeter

Vacuum-Tube Voltmeter (VTVM) TS–375/U is a typical general-purpose high-impedance a-c and d-c voltmeter for servicing and testing electronic equipment. It is intended particularly for voltage measurements where the sensitivity or frequency range of standard voltmeters is insufficient, for example, for measuring d-c grid bias voltage, audiofrequency voltage or radiofrequency voltage.

The input impedance is sufficiently high to avoid consideration of the instrument current drain upon the circuit undergoing measurement. For example, in circuits where the current drain of a 20,000-ohms per volt voltmeter would cause a serious error or operational upset to the circuit, this instrument can be used without difficulty.

For a-c measurements the frequency range includes the radiofrequency spectrum up to approximately 300 megacycles, whereas standard a-c voltmeters have a useful upper frequency limit of a few thousand c.p.s.

The instrument includes an a-c probe which contains a diode rectifier tube and d-c blocking capacitor arranged to provide a minimum frequency error at high frequencies. For the low-frequency panel terminal connections a larger internal blocking capacitor is used. It also includes a d-c probe which contains an isolating resistor (5 megohms) to minimize disturbance to circuits containing a.c. in which the d-c voltage is being measured.

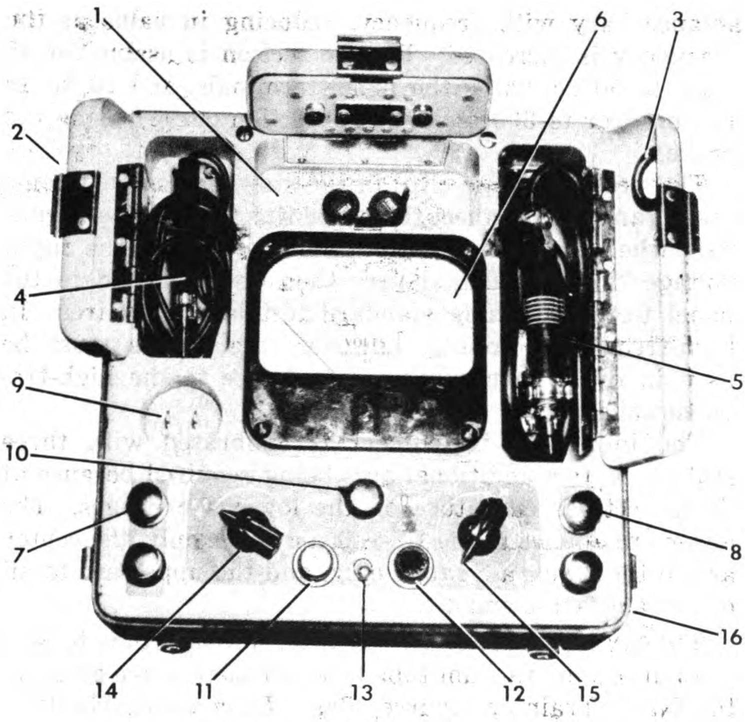


Figure 11–14.—Voltmeter TS–375/U, front panel view.

1. Accessory stowage compartment.
2. Leadout grommet for d-c probe.
3. Leadout grommet for a-c probe.
4. D-c probe.
5. A-c probe.
6. Indicating instrument.
7. D-c panel terminals.
8. A-c panel terminals.
9. Power cable receptacle.
10. Zero adjuster.
11. Standby heater pilot lamp.
12. Line pilot lamp.
13. Line switch.
14. Selector switch.
15. Range switch.
16. Rubber gasket.

This instrument requires a power source of 105 to 125 volts, 50 to 1,600 cycles. It provides the following d-c ranges, positive or negative polarity—1.2, 3, 12, 30, 120, and 300 volts—with a constant input resistance of 30 megohms. The a-c ranges are 1.2, 3, 12, 30, and 120 volts with a 5-megohm resistance on all ranges shunted by approximately 5 μf. when using the probe externally. The effective shunt resistance and shunt capacitive reactance vary with frequency, reducing in value as the frequency is increased. The a-c section is usable for 40 c.p.s. to 50 kc. using the panel terminals, and 10 kc. to 150 mc. (or to 300 mc. with correction curve) using the probe.

The probe stowage clips in the stowage compartments are arranged to connect the probe tips to the panel terminals when the probes are correctly stowed. The input voltage to be measured may then be connected to the panel terminals using standard test leads or wires. In high-frequency circuits, however, the probes must be used in order to minimize disturbance to the high-frequency circuits.

The indicating instrument is calibrated with three scale arcs, two additional arcs being required because of diode rectifier curvature on the lower a-c ranges. The lower arc applies to the 1.2-volt a-c range only, the center arc to the 3-volt a-c range only, and the upper arc to all other ranges, a-c and d-c.

THEORY OF OPERATION.—A brief introduction to and the purpose of vacuum-tube type voltmeters are given in the Navy Training Course, *Basic Electroncis*, NavPers 10087, chapter 13.

Since the TS–375/U voltmeter is widely used in aviation electronics maintenance, its theory of operation will be covered somewhat in detail.

This test set may be divided into six major circuit sections as shown in the block diagram of figure 11–15. Energy from the circuit under test is picked up through either probe, is subdivided in the switch section, and is fed to the input amplifier grid. The power supply energizes the amplifier which, in turn, operates the indicating instrument in proportion to the potential at the amplifier input grid.

The POWER SUPPLY is a conventional assembly and supplies the necessary voltages and currents to the amplifier section.

The D-C AMPLIFIER is a compound feedback circuit com-

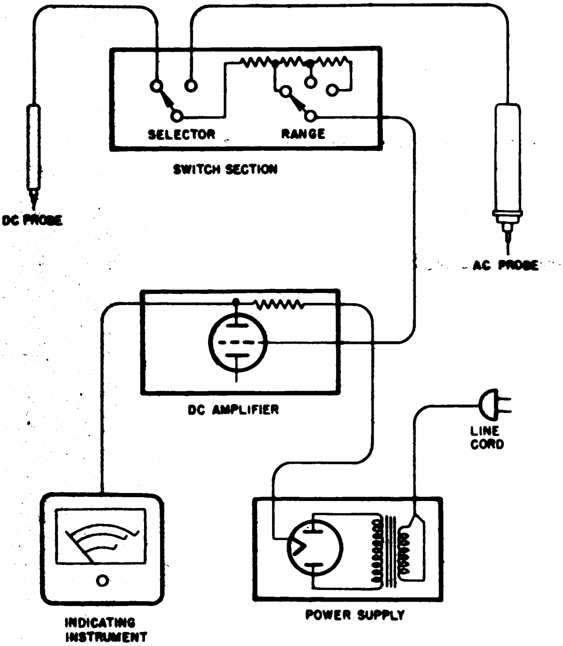


Figure 11–15.—Block diagram—TS–375/U.

bining regeneration and degeneration to obtain gain stability. Whenever gain stability is desired, it is general amplifier practice to employ an amplifier having a gain much larger than the final gain desired, and degenerating the excess gain by reverse feedback from the output back to the input end. In theory the higher the gain inherent in the amplifier and the greater the resultant degeneration, the greater the overall stability in the face of changes in the gain ratio of the amplifier proper. Thus, stability is in part a problem of obtaining a high gain so that the advantages of a high degree of degeneration can be realized.

In the compound feedback circuit the additional gain necessary for stability is obtained by regeneration. By this method a gain ratio far greater than the actual gain of the amplifier tubes proper can be realized.

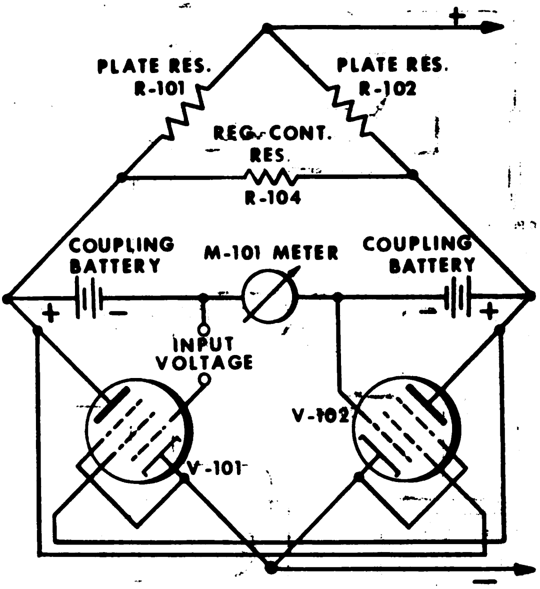


Figure 11–16.—D-c amplifier, simplified circuit.

The action of the amplifier is illustrated by the simplified functional diagram of figure 11–16. The circuit is essentially a bridge with the two amplifier tubes and the two plate resistors forming the four bridge arms. The d-c input voltage is impressed upon the control grid of *V*101 while *V*102 serves as a compensating tube to minimize supply voltage variations and drift effects upon the bridge balance. The regenerative action is obtained by cross-connecting the plates and screen grids of the tubes in the manner of a multivibrator type of oscillator. The screen grids then serve as secondary control electrodes rather than as screen grids in the conventional sense. The cross-connected resistor *R*104 serves as a control of the degree of regeneration obtained by loading the tube plate circuits. The cathode-connected suppressor grids serve to suppress secondary emission effects in the conventional manner.

Degeneration is obtained by conductively coupling the plate of each tube to its control grid through coupling batteries. The grid of $V101$ is coupled through the input circuit whereas the control grid of $V102$ is directly connected to the negative end of its coupling battery. The output indicating voltmeter is connected between the negative ends of the coupling batteries so that bridge unbalance in response to an input voltage will supply an unbalance current to the voltmeter circuit for indication.

The voltage appearing across the output voltmeter in response to an input voltage has a polarity in opposition to the input polarity, as indicated by the polarity signs on the input and indicating instrument circuits. Thus, the voltage appearing between the two control grids is the difference voltage between input and output and must be considerably smaller than the total input voltage swing. Furthermore, as the tube gain is increased by regeneration in the plate-screen grid circuit, the difference becomes less and less to the point where the output voltage is equal to the input voltage. The regeneration control resistor $R104$ is adjusted to this condition which is the optimum point for stability.

At this point of optimum adjustment the amplifier plate-screen grid circuit is critically regenerative, and would be on the verge of self-sustained oscillation if the degenerative connections to the control grids were disconnected or otherwise rendered inoperative. In the overall sense the amplifier is degenerative but internally consists of a regenerated section around which a degenerative circuit is connected. The amplifier may also be visualized as an amplifier having an infinite gain (secured in this case by critical regeneration) which is completely degenerated externally; thus, the stability improvement envisioned in any degenerative amplifier by greater gain and more degeneration is carried to its logical endpoint.

At the point of critical regeneration the screen grids do the actual work of unbalancing the bridge to produce an output while the control grids exert only a transient type of control to initiate the action. For any steady value of input voltage the potential of each control grid with respect to its cathode is the same, and the potential difference between the grids is zero. This feature of zero grid voltage excursion in response to an input voltage is highly desirable because it entirely removes the effect of the curved grid voltage-plate current (e_g-e_p) characteristic common to all tubes, and the amplifier is strictly linear in its input to output relationship.

The practical amplifier circuit is shown in figure 11–17. It is entirely similar in function to the simplified circuit shown in figure 11–16, but the addition of certain components is dictated by practical circuit limitations and operating conditions.

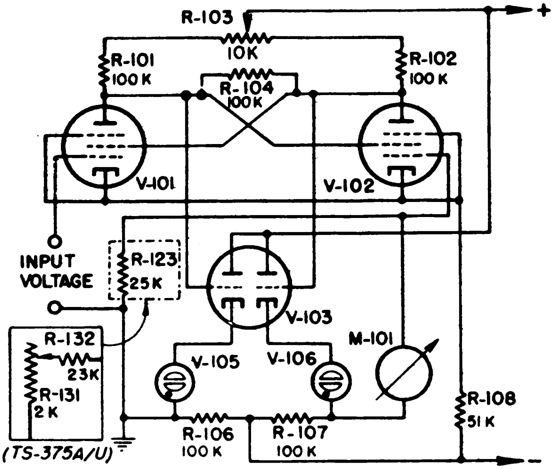


Figure 11–17.—D-c amplifier, practical circuit.

Voltage regulator tubes *V*105 and *V*106 replace the coupling batteries. The low variational resistance characteristic to this type of regulator tube is similar in effect to a battery. The voltage regulator tubes require a continuous keep-alive current, necessitating the addition of keep-alive bleeder resistors *R*106 and *R*107, from each tube to the negative plate supply lead. The keep-alive current required by the coupling tubes would be an excessive current drain from the amplifier tube plate circuit proper so a twin-triode tube is added as a cathode follower between the amplifier plates and the loaded coupling tube circuits. No further amplification is contributed by this tube except for the advantages gained by removing all plate loading from the amplifier tubes proper.

Because of inherent differences between tubes and other circuit components, it is necessary to initially balance the bridge to zero output with zero applied input. The zero adjusting potentiometer *R*103 serves this purpose by controlling the plate series resistance of one amplifier tube relative to the other.

A cathode resistor is added between the amplifier tube cathodes and the negative plate supply lead because of the voltage drop added by the bleeder resistors.

The SWITCH SECTION consists of a range switch *S*102 for selecting the desired range of input voltage, and a selector switch *S*103 for selecting the desired function, d-c input of either polarity or a-c input. Refer to figure 11–18, which is a practical wiring diagram of the voltmeter and will be used throughout the remainder of this discussion.

The range switch with its connected resistors is essentially an input attenuator and changes range by accurately dividing the input voltage. In addition, on a-c ranges it is necessary to correct for the contact potential of the diode rectifier in the a-c probe. In thermionic (hot cathode) rectifiers the cathode emission alone will develop a negative potential on the plate with no a-c input. This is termed the "contact potential" of the plate with respect to the cathode. The probe diode contact potential is compensated in the instrument by adding a similar diode to the grid circuit of the other amplifier tube.

The contact potential of the balancing diode is applied through a second attenuator section similar to the range

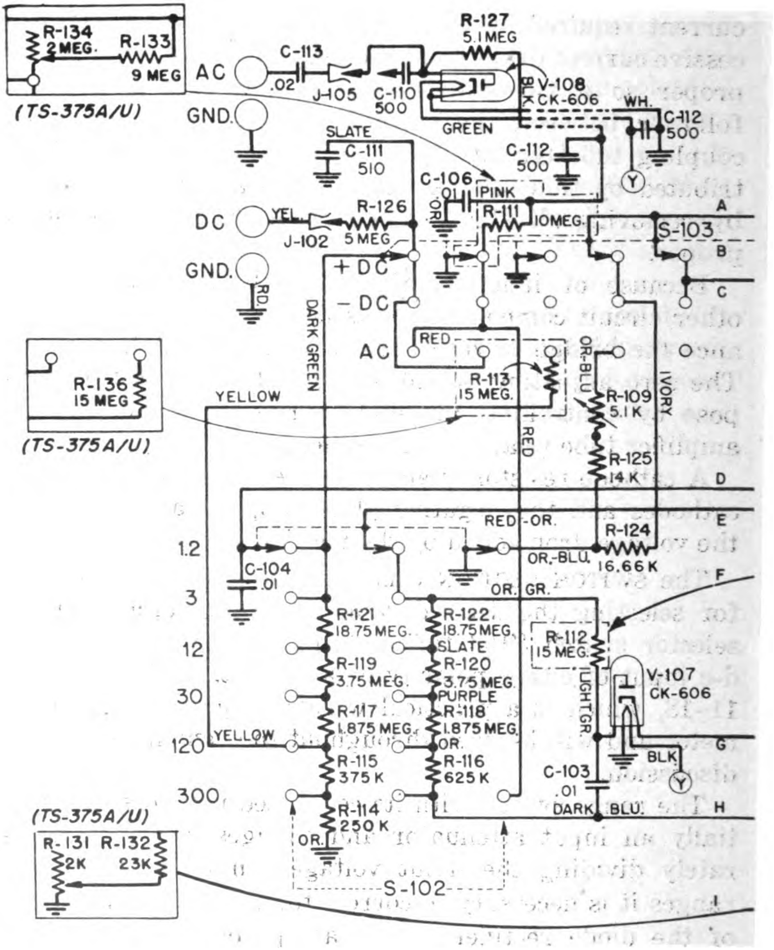


Figure 11–18.—Voltmeter TS–375/U, practical wiring diagram.

attenuator. The attenuation is not changed when switching from the 3-volt to the 1.2-volt range; instead the range is changed by changing the range of the output voltmeter by the application of a shunting resistor *R*124 across the voltmeter resistor. Thus, when switching from the 3-volt to the 1.2-volt range, the only connection change is reduction of the output voltmeter range by shunting. The range switch also grounds the a-c probe

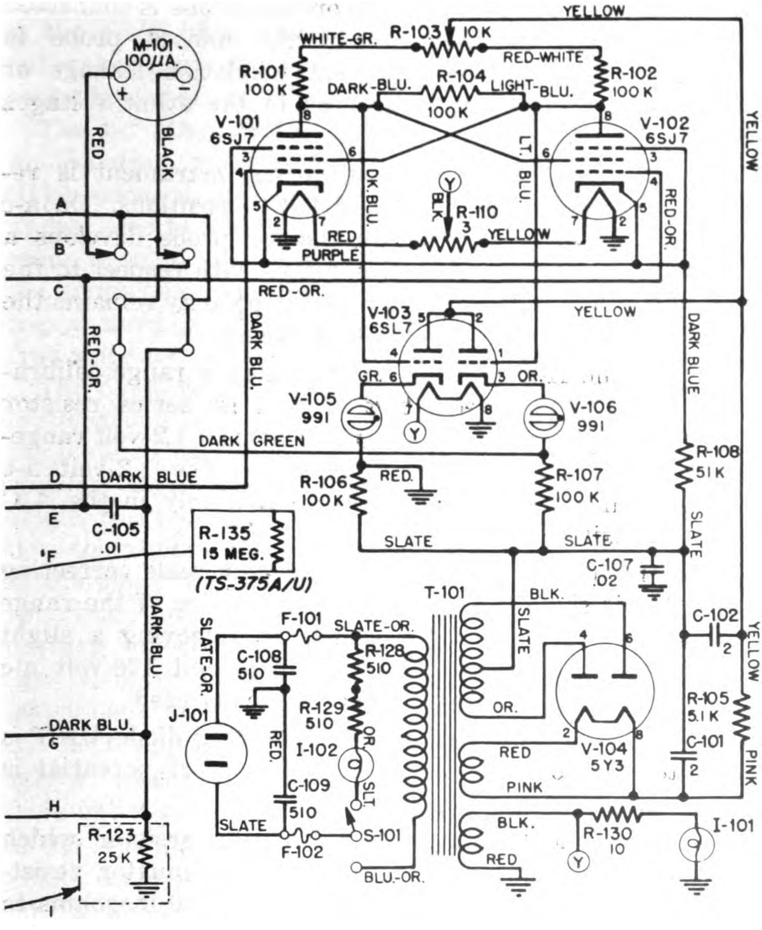


Figure 11–18.—Voltmeter TS–375/U, practical wiring diagram. Continued.

lead in the 300-volt position so that the instrument will not be functional on a-c input voltages that can damage the diode rectifier.

The selector switch has three positions, D-C +, D-C—, and A-C, and primarily serves to switch the range switch between the d-c and a-c probe and to reverse the polarity of the indicating instrument. It also performs the following functions:

1. In each position the appropriate probe is connected to the range attenuator, and the unused probe is grounded to completely prevent insulation leakage or capacitive coupling interferences in the event voltages are applied to the unused probe.

2. The polarity of the indicating instrument is reversed between the D-C+ and D-C— positions. On a-c ranges the diode rectifier in the a-c probe develops a rectified d-c voltage that is negative with respect to the ground so the indicating instrument polarity remains the same between the D-C and A-C positions.

3. In the A-C position the 1.2-volt a-c range calibration adjustment rheostat *R*109 and its series resistor *R*125 are connected in place of the normal 1.2-volt range-changing resistor *R*124. This makes the 1.2-volt a-c range calibration adjustment effective only in the A-C switch position.

4. In the A-C position a 15-megohm scale correcting resistor is shunted across the lower section of the range attenuator. This resistor assists in removing a slight scale tracking error on the 12-, 30-, and 120-volt a-c ranges where individual scale arcs are not used.

5. In both d-c positions the balancing diode *V*107 is shorted out because the balancing contact potential is only required on a-c ranges.

The D-C PROBE contains an isolating resistor which together with the 25-megohm range attenuator resistance presents a total input resistance of 30 megohms to the circuit in which a measurement is being made. This resistor serves to minimize capacitive loading of the measured circuit which is important if the circuit contains RF voltages, and is particularly important in resonant circuits which might be thrown out of alinement during the measurement.

When making measurements in purely d-c circuits, the d-c probe may be stowed in the d-c probe compartment and the voltage for measurement applied to the panel input terminals by means of standard test leads or wire connections. When the probe is clipped into its holder in the probe compartment, the ungrounded panel terminal makes connection to the probe tip through the lower probe holder clip.

The **A-C PROBE** consists of a diode rectifier tube *V*108, an isolating resistor *R*127, and a blocking capacitor *C*110, encased in a cylindrical bakelite case similar in appearance to a standard test probe but larger in diameter. In addition to the conventional test probe-tip prod the contact end carries a circumferential ring for making a short ground connection when measuring high-frequency a-c voltages.

The diode rectifier tube is shunt-connected across the input circuit with the cathode connected to ground and the plate connected to the ungrounded input lead. It thus rectifies the a-c input by passing the positive half cycles of a.c. to ground. A few half cycles of positive conduction to ground then will build up a negative potential at the plate. The blocking capacitor will then in a short time build up a charge above ground almost equal to the positive half-cycle peaks of the a-c wave. This d-c voltage is delivered through the isolating resistor to the instrument proper for indication. After the a-c input has been applied for several cycles, the diode conducts only a sufficient portion of each positive peak to supply the small d-c current demanded by the range attenuator in the instrument.

The d-c voltage developed at the diode plate is therefore almost equal to the positive peaks of the input wave, and the instrument is essentially a peak, or crest, voltmeter. However, the reading should be in terms of the effective a-c voltage, or root mean square (r.m.s.) value, which is lower than the peak value. The instrument is therefore calibrated in terms of the r.m.s. value of a pure sine waveform, which has a peak value equal to $\sqrt{2}$ (or 1.414) times its r.m.s. value. Thus, the rectified d-c voltage developed at the diode plate is approximately 1.414 times the indicated a-c voltage. Conversely the r.m.s. value of a sine wave is 1/1.414 (or 0.707) times the peak value.

The excess d-c voltage developed by the a-c probe is compensated for by an additional resistor *R*111 in the a-c probe-attenuator lead.

At high levels of a-c input to the probe the rectified d-c output is almost directly proportional to the a-c input. On low ranges, however, the nonlinear space path resistance, characteristic to all rectifiers, becomes increasingly effective and calibration will not fit a uniform meter scale with sufficient accuracy. The low a-c ranges, 1.2 and 3 volts, therefore, have individually calibrated scale arcs.

An important feature of the a-c section is the frequency range over which it is usable. The low-frequency limit is determined by the impedance of the input blocking capacitor which becomes appreciable and causes a voltage drop and a reading error. This indicates use of a large value of blocking capacity for low-frequency accuracy. On the other hand, a physically large blocking capacitor would increase the inductance and capacitance to ground in the input circuit lowering the upper useful frequency limit. Thus a large capacitor is desirable at low frequencies and a small capacitor is desirable at high frequencies. This two-capacitor arrangement is realized by equipping the probe proper with a small capacitor *C*110, 500 $\mu\mu$f., and arranging the probe stowage holder to contact the probe after the small probe capacitor so that a larger blocking capacitor *C*113, 0.02 μf., can be located in the internal lead to the a-c panel terminal. Thus by using the a-c probe externally for high frequencies and the panel terminals for low frequencies a greater frequency range is obtained than would be the case with a single blocking capacitor in the probe only. The small coaxial contact ring immediately behind the probe-tip prod serves to make this connection when the probe is stowed in its holder.

In high-frequency measurements the ground return connection must be short and in close proximity to the ungrounded connection. For example, if the ground connection were made at the ground panel terminal at a frequency of 50 megacycles, the entire circuit loop comprising the probe cable and the ground lead may resonate and cause readings that are high by as much as several hundred percent. Conversely, this type of connection is perfectly good practice at, for example, 500 kc. The a-c probe is equipped with a close ground connection point in the form of a coaxial ring for use at high frequencies. This ground ring also serves as a convenient point of attachment for accessory connection hardware such as the ground clip prod.

The INDICATING INSTRUMENT *M*101 is a microammeter having a current range of 0–100 microamperes and an internal resistance of approximately 1,000 ohms. It indicates the amplifier output voltage by responding to the current that flows through the output resistor. By placement in the current side of the output circuit rather than directly in series with the output resistor, its internal resistance does not figure in the calibrated circuit, and need not be adjusted accurately.

OPERATION AND ADJUSTMENT.—When power is applied to the instrument, the white (heater) pilot should light with the line switch "OFF." This indicates that power is applied to the electrical heater included within the instrument to reduce the internal relative humidity when not in use. When the line switch is turned on, the red pilot light will light and the heater and its pilot will be deenergized. After a 30-second warmup period, short circuit the input terminals and zero the instrument with the panel ZERO ADJUST potentiometer. Shorting the input circuit during zero adjustment is recommended to avoid the influence of stray leakage or induced voltages which might otherwise lead to setting to a false zero. This is particularly important on low a-c ranges. The zero adjustment is not affected between d-c ranges and no readjustment should be necessary between d-c ranges. Between a-c ranges, however, a slight zero shift may take place, and for accurate measurements zero readjustment on each a-c range is recommended.

NOTE: One input terminal of both a-c and d-c input circuits is grounded to the chassis of the instrument. Consequently, the voltmeter is adaptable only to measurements between circuit points one side of which is grounded, or which may be connected to the voltmeter chassis safely and without disturbance to the circuit.

D-C VOLTAGE MEASUREMENTS of either polarity may be measured by setting the selector switch to the d-c position indicating the polarity to be applied to the ungrounded input terminal. For measurements in circuits containing radiofrequency voltage use the d-c probe for connection to the ungrounded side; in straight d-c circuits either the panel input terminal or the probe may be used. In either case the ground connection is made to the ground panel terminal.

The following statements relate to A-C MEASUREMENTS:

1. For measurements at frequencies lower than 10 kc. the panel terminals must be used.

2. For measurements at frequencies higher than 50 kc. the probe must be used.

3. For measurements from 10 kc. to 50 kc. either the probe or the panel terminals may be used.

4. Below one megacycle the ground connection may be made to the panel ground terminal or to the probe ground ring.

5. Above one megacycle the ground connection must be made to the probe ground ring, and kept as short in length as feasible.

6. In general, as the frequency increases toward the top limit of 300 megacycles, the input lead lengths, both ground and ungrounded sides, must be made shorter. For example, at 40 megacycles leads three or four inches long are permissible, but above 100 megacycles the prod tips proper must contact the circuit directly. Thus at the higher frequencies it is not possible to measure volt-

age between circuit points that are widely separated. The ground ring spring clip can serve as a ground prod up to 100 megacycles, but at higher frequencies direct contacting to the ground ring is recommended. The alligator spring clip may be used on the ungrounded prod at frequencies not exceeding 50 megacyles.

7. On frequencies above 100 megacycles, the loading on the circuit under test may be reduced by removing the probe tip.

8. Any method of connection permissible to a certain maximum frequency may be used at any lower frequency down to the bottom frequency limit of 10 cycles, except that the panel terminals must be used below 10 kc.

9. The readings are essentially without frequency error from 50 cycles to 150 megacycles. Above and below these frequencies correction factors as indicated by the curves included in the *Handbook of Service Instructions* may be applied to extend the frequency range from 10 c.p.s. to 300 mc.

The ACCURACY of the instrument is summed up as follows:

Section	*Percent of full scale range*
D-c, all ranges	3
A-c, all ranges:	
10 to 50 c.p.s.	5, with correction curve
50 c.p.s. to 50 mc.	4, without correction
50 mc. to 150 mc.	6, without correction
50 mc. to 150 mc.	3, with correction curve
150 mc. to 300 mc.	8, with correction curve

(NOTE: The diode rectifier responds to the positive peak value of the applied voltage whereas the indicating instrument is calibrated in r.m.s. value of a pure sine wave. Thus if the applied waveform is other than a pure sine wave, an error may be expected in the indicated voltmeter reading.) With distorted waveforms the reading may be interpreted as 0.707 times the positive peak of the voltage being measured.

With the worst possible phase distribution of harmonic components in the applied waveform, the maximum error cannot exceed the sum of the percentages of all the harmonics. For example, if the total harmonic content is known to be less than 10 percent, the waveform error can by anywhere between 0 and 10 percent. The error may either increase or decrease the reading.

When MODULATED WAVEFORMS are applied, the diode rectifier will respond to the recurrent modulation peaks. The reading will thus be high by an amount dependent upon the percentage modulation and the waveform of the modulation envelope. This error may be considered approximately equal to the percentage modulation if the modulation envelope is a sine wave. With pulse modulation another error becomes apparent if the duty cycle is very short, as in the case of radar modulation. The duration of the peaks becomes so short relative to the fundamental pulsing frequency that the rectified voltage drops appreciably between pulses. This causes a low reading error that becomes larger as the duty cycle is made smaller.

Occasionally the effect of TRANSIENT SURGES from power line switches or other sparking devices will be experienced. They are evidenced by erratic motions of the indicating instrument. If sufficiently strong to project through the internal bypassing, they may be injected through the line connection to the instrument; but generally they are introduced through inductance in the ground return circuit mutual to the input and the path of the surge.

The following corrective measures are suggested:

1. Locate the source of the disturbance and, if possible, shut down the offending device during measurements. Such devices are potential radio noise generators and should be corrected as soon as discovered.

2. Ground the instrument chassis by physically grounding one of the input ground terminals.

3. Shorten the ground lead to the measured circuit as much as possible.

4. Place line filters in the power cord to the instrument and, if possible, in the power connection to the device being measured.

Transient surges will have no effect unless the surge peak is larger than the peak value of the input voltage being measured. Their effect will therefore generally be noticed only with low or zero applied input voltages.

The instrument may be OVERLOADED by applying an excess voltage of the type (d-c or a-c) being measured, which will drive the indicating instrument off scale, or by applying an excess voltage of opposite type, which will not indicate on the instrument but may cause internal damage.

Overloading on the d-c ranges can result in burnout of the d-c probe resistor. Overloading on the a-c ranges can result in burnout of the a-c probe resistor, flashover in the probe blocking capacitor, or destruction of the diode rectifier.

MAINTENANCE.—The most important aspect of ROUTINE PERFORMANCE CHECKING is testing for zero adjustment. Check the instrument thoroughly by applying d-c and a-c voltages of known values to the various ranges. Check the panel zero adjustment on each range before each calibration check. If the instrument pointer cannot be set to zero on any of the ranges, or erratic readings are noted, proceed with the necessary adjustment or repair.

Do not attempt to operate the instrument outside of its case near high-frequency power equipment. RF voltages may be induced into the circuit in sufficient amount to cause erratic operation.

Detailed information dealing with proper maintenance procedures is given in the *Handbook of Service Instructions* for the voltmeter.

SPECIAL MEASUREMENTS.—Voltmeter TS–375/U can be used to measure electrical quantities other than voltage by external correction of relatively simple accessory components. It can thus serve as an emergency measuring device whenever special test equipment normally used for this purpose is not available. The high input impedance compared with conventional d-c and a-c voltmeters allows a materially greater range and sensitivity in such special measurements than could be obtained with the same circuits using conventional voltmeters. It may be used for measuring megohm resistances, capacitance, inductance, low a-c voltages, low d-c currents, and decibels. The necessary charts and information for performing these measurements are contained in the *Handbook of Service Instructions.*

QUIZ

1. Any variation from the normal time between periodic inspections will be found in the
    a. *Handbook of Service Instructions*
    b. *Handbook of Maintenance Instructions*
    c. *Handbook of Aircraft Inspection Requirements*
    d. *Handbook of Overhaul Instructions*

2. Minimum performance standards for electronics equipment will be found in which section of the *Handbook of Service Instructions?*
    a. Section I
    b. Section V
    c. Section IV
    d. Section III

3. Detailed instructions for removal and replacement of components will be found in which section of the *Handbook of Service Instructions?*
    a. Section V
    b. Section III
    c. Section VII
    d. Section VI

4. When directed to ship electronics equipment, information in regard to proper packing would be found in
   a. Section III of the *Handbook of Service Instructions*
   b. Section VI of the *Handbook of Service Instructions*
   c. the *Handbook of Overhaul Instructions*
   d. *Illustrated Parts Breakdowns*

5. What is the meaning of the term CALENDAR INSPECTION?
   a. A regularly scheduled periodic inspection
   b. An annual inspection
   c. Inspections which take the place of periodic inspections if a required number of flight hours are not accomplished in a specified length of time
   d. Inspections which must be carried out on a certain specified day of each month

6. Who writes the *Handbook of Aircraft Inspection Requirements?*
   a. ComAirLant
   b. BuAer
   c. OpNav
   d. Aircraft manufacturer

7. Climatic deterioration is most prevalent in
   a. temperate zones
   b. the Arctic
   c. the Antarctic
   d. tropical zones

8. What is the main cause of malfunctioning of electronics equipment at high altitudes?
   a. Lack of oxygen
   b. Low temperatures
   c. Lower breakdown voltage
   d. Condensation

9. Detailed information on the suppression of radio noise in aircraft can be found in
   a. T.O. No. 16–1–45
   b. NavAer 16–1–518
   c. T.O. No. 16–1–44
   d. NavPers 10087

10. An efficient and systematic plan of troubleshooting can best be accomplished by
    a. use of proper test equipment
    b. reference to the *Handbook of Service Instructions*
    c. a good understanding of the theory of operation of the equipment
    d. reference to the *Handbook of Aircraft Inspection Requirements*

11. Static discharge wicks are used to
    a. combat manmade noises in aircraft communications equipment
    b. prevent development of large static charges on an aircraft when flying through areas of precipitation
    c. overcome effects of high magnetic fields
    d. reduce the effects of atmospheric static on VHF and UHF equipment

12. Which of the following is not used for suppression of manmade radio noise?
    a. Insulated antenna system
    b. Isolation
    c. Shielding
    d. Filtering

13. With all safety devices operating correctly, what is the most likely trouble source in high altitude aircraft electronics equipments?
    a. High voltage arc-over
    b. Excessive brush wear
    c. Switch contacts "welding"
    d. Waveguides exploding from excess pressure

14. Small targets are most likely to be lost in radar reception due to
    a. weak transmitter output
    b. low PRF
    c. low receiver sensitivity
    d. intermittent modulator trigger

15. The Test Set TS–147A/UP is designed to be used with radar operating in the
    a. S band
    b. K band
    c. P band
    d. X band

16. Moving the signal width control on the TS–147A/UP will
    a. vary the range of frequency modulation of the RF generator
    b. vary the repeller plate voltage of the RF generator and change modes
    c. vary the time of the sawtooth applied to the RF generator
    d. have no effect on the center frequencies generated by the RF generator

17. There are two thermistors used in the circuit of the TS–147A/UP. They are a
    a. bead thermistor with a negative temperature coefficient of resistance as one leg of the bridge and a disk thermistor with a negative temperature coefficient of resistance in series with the meter, for temperature compensation
    b. disk thermistor with a positive temperature coefficient of resistance as one leg of the bridge and a bead thermistor with a negative temperature coefficient of resistance in series with the meter, for temperature compensation
    c. bead thermistor with a negative temperature coefficient of resistance as one leg of the bridge and a disk thermistor with a positive temperature coefficient of resistance in series with the meter, for temperature compensation
    d. disk thermistor with a negative coefficient of resistance as one leg of the bridge and a bead thermistor with a positive coefficient of resistance in series with the meter, for temperature compensation

18. When using the Test Set TS–147A/UP to check the performance of radar, which of the following connections to the test set is unnecessary if the radar transmitter is operating?
    a. Power to the test set
    b. RF coupling to the test set from the radar
    c. Trigger to the test set from one of the radar trigger outputs
    d. Connection from the radar video output to an oscilloscope

19. The 600-volt probe supplied with the Oscilloscope AN/USM–24 has an attenuation factor of
    a. 4:1
    b. 10:1
    c. 1:1
    d. 5:1

20. What is the smallest signal which can be observed on the AN/USM–24 oscilloscope?
    a. 1.00 volts
    b. 2.00 volts
    c. 0.10 volt
    d. 0.01 volt

21. When using the AN/USM–24 oscilloscope to observe the time relationship between two wave forms, the oscilloscope should be triggered by
    a. the incoming signal
    b. internal trigger generator
    c. any external trigger source
    d. the same trigger being applied to the circuits to be compared

22. Which characteristic below applies to the vacuum-tube voltmeter TS–375/U?
    a. A-c voltage measurements are possible at all frequencies under 1,000 mc,
    b. Voltage division for the high voltage scales is provided by use of isolating resistors in the a-c and d-c probes
    c. The input resistance of the meter is 20,000 ohms/volt
    d. Gain is stabilized by utilizing a positive and a negative feedback circuit

23. For accurate measurement of grid bias of an amplifier tube, with a TS–375/U vacuum-tube voltmeter, measurements are made from
    a. grid to cathode
    b. grid to ground
    c. cathode to ground
    d. grid to ground and cathode to ground; then compute the difference

24. Which of the following measurements would not normally be taken with the TS–375/U vacuum-tube voltmeter?
    a. Grid bias voltage on an amplifier tube
    b. B+ voltage on the power amplifier stage of a communications transmitter
    c. RF voltage present in an RF amplifier stage
    d. Line voltage applied to electronics equipment

CHAPTER 12

AVIATION ELECTRONICS SUPPLY

YOUR JOB IN SUPPLY

It is important that all naval personnel have a general knowledge of the principles of our present supply system in order to fully and correctly utilize the system. It is of primary importance that administrative and supervisory personnel have a working knowledge of the methods used to obtain and properly account for their particular supplies.

This chapter outlines your functions with regard to supply and tells you in a general way what you should know and be able to do in fulfilling this part of your job. As a first class or chief, you will probably be either fully or partially in charge of maintenance and operation in your division. You will, therefore, be responsible for seeing that an adequate supply of spare parts, tubes, and other materials are available, and to help in planning for future needs. These, in general, are your main responsibilities with regard to supply.

AVIATION SUPPLY ORGANIZATION

The Bureau of Supplies and Accounts (BuSandA) is the Navy's organization for determining policies and outlining the procedure to be followed in performing supply and fiscal functions both afloat and ashore. Specific and detailed information on Navy supply is contained in the *Bureau of Supplies and Accounts Manual.*

The Aviation Supply Office (ASO) is a joint agency of the Bureau of Supplies and Accounts and the Bureau of

Aeronautics, and is charged with the exercise of inventory control for aviation material. It does not stock materials but serves as the control center.

The primary functions of ASO are as follows:

1. Procurement of most aeronautical supplies from manufacturers or other government departments.
2. Disposal of surplus materials.
3. Distribution, storage, and inventory control of aeronautical materials after procurement.
4. Completion of statistical information to aid in future procurement and distribution of aeronautical materials.
5. Classification and cataloging of aeronautical materials, and the distribution of such materials.
6. General control and rationing of critical aeronautical materials with the exception of that material controlled directly by BuAer and fleet commands.

NOTE: The ASO controls most of the aviation electronics equipment; however, some components, such as electron tubes, are under the control of the Electronics Supply Office (ESO). The trend is to transfer all common parts to the ESO.

CATALOG OF NAVY MATERIAL

To order material and to assure receipt of it by your division, you make out a REQUISITION. But, to be sure you receive the material, you have to do more than list on a requisition form the names of the items you want.

Because of the great variety of different items carried in the Navy supply system, it is necessary to have a workable method of keeping order and control over procurement, inspection, storage, transportation, issuing of, and accounting for supplies.

The *Aviation Supply Office Catalog*, a section of the *Catalog of Navy Material*, is the basic stock catalog used by aviation supply personnel. This ASO catalog contains the answers to the majority of the material problems in aviation supply; and because of this, a rather complete knowledge of its contents and use will prove of untold value to the electronics technician. Other sections of the *Catalog of Navy Material* that you should become familiar with are BuShips Sections, Part II (Electronic) and Part III (Electronic Equipment), and the General Stores Section.

By spending some of your spare time in becoming familiar with these catalogs and other supply publications, you will be in a position to expedite parts requisitions when time may be at a premium. An excellent description of this catalog may be found in the Navy Training Course, *Aviation Storekeeper,* Volume 1, NavPers 10396–A, chapter 12.

PROCUREMENT

A high degree of aircraft availability is impossible to maintain unless there is an adequate supply of spare parts, tubes, and materials. The maintenance spares and parts your activity will be permitted to carry on hand will be governed by the classification given the activity by the Bureau of Aeronautics. The amount of material provided is determined by the BuAer *Initial Outfitting Lists* and *Allowance Lists* (the term allowance lists generally refers to both). These lists are used at the time of initial commissioning or reactivation of a unit. They are lists of equipment and material determined from known or estimated requirements as necessary to place and maintain aeronautical activities in a material readiness condition.

INITIAL OUTFITTING LISTS.—An initial outfitting list is essentially a "supply type" list, that is, a listing of spare parts and consumables required at initial outfitting. Thereafter, the range and quantities of these items are to be as dictated by local usage and requirements. However, these lists may also be used by supporting supply departments as a guide in determining stock levels when supporting a given number of aircraft or equipments.

The section "R" initial outfitting lists are of major interest to electronics personnel, and are as follows:

1. Standard Aeronautical Electronics Material Section "R" (NavAer 00–35QR–4). This list contains common resistors, capacitors, and other parts that are used in three or more different equipments. For example, if a particular resistor that is used in the AN/ARC–27 is also used in two other equipments, it is deleted from the Section "R" for the AN/ARC–27 and carried in this section. It is worthwhile to understand the above allowance list, for should an item not be found in the allowance list for a particular type of equipment, it will most likely be found in the NavAer 00–35QR–4 allowance list.
2. Aeronautical electronic accessories are listed in BuAer Allowance List Section "R" (NavAer 00–35QR–6). This list contains accessories that have common usage with many different types of electronic equipment, such as microphones, headsets, and insulators.
3. Electronic tubes are listed in NavAer 00–35QR–7. This is a listing of equipments requiring electron tube support. The equipments are arranged alphabetically and numerically. The type, stock number, quantity of each type used, and tube life hour code for each tube used by that equipment is given.

   Tube life hour code is a code which assigns each electron tube to a group or family which experience has shown may be expected to require replacement after approximately the same number of hours in service. Each code represents a different service life span.

Following the alphabetical listing is a chart designating the Electron Tube Allowances by percentage for the number of tubes supported, by tube life hour code.

This allowance list is very useful in determining the types and quantities of tubes required for squadron deployment on a self-supporting basis.

4. The NavAer 00–35QR–30 Series lists peculiar maintenance parts, including spare major components and subassemblies, required for the maintenance support of the concerned electronic equipment.

Because Initial Outfitting Lists identify supply stock items they are also valuable as listing of material that is consumed but not stocked by the operating squadron or FASRon. The lists have been made available to squadrons and all Fleet activities to be used as a ready reference when ordering from the stocks of the supporting supply activity. Materials represented by these lists are available to Navy squadrons through "Ready-Issue Stores."

ALLOWANCE LISTS.—An allowance list is essentially an "equipment type" or "requirement type" list. In electronics maintenance the only requirement type lists are the NavAer 00–35QR–5 for the test equipment, the Section G for shop, handling and servicing equipment, the Section K for naval aeronautic publications and forms, and the Section U for hand tools. The allowance list should be considered as a necessary instrument in material relations between the shore establishment and the needing unit or activity.

The squadron or FASRon electronics officer is directly concerned with allowance lists covering shop and maintenance (operational) equipment in order to determine that all items required for a satisfactory state of material readiness are on hand for use.

ACCOUNTABILITY SYMBOLS.—The responsibility for accountability in connection with electronic equipment is a serious business. All items received will be identified with a symbol that determines the type of accountability required. These symbols are given in the Allowance and Initial Outfitting Lists for each item listed. The meaning of the symbols may also be obtained from any of the NavAer 00–35QR publications.

INVENTORY CONTROL.—Inventory control may be defined as "the formulation and administration of a system of policies and directives concerned with the determination of requirements or excesses and the action necessary to satisfy such requirements or dispose of such excesses." Examples of the activities that perform these functions, under overall BuSandA direction, are the Aviation Supply Office (Philadelphia), the Electronics Supply Office (Great Lakes), and the Yards and Docks Supply Office (Port Hueneme).

Within the inventory control operation at established activities of the shore establishment, Ready-Issue Stores, and Shop Stores are intended to be what might be called the little corner grocery store or retail outlet intended to get the spare parts and consumables as close to the user as possible and still keep the materials in the accounts of the Navy until they are really needed.

The FASRon's have the responsibility for establishing and operating the ready-issue stores under a Supply Corps officer of the shore activity. The range and quantity of items carried is determined by the supply officer, guided by usage data, applicable Allowance Lists and Initial Outfitting Lists, and recommendations submitted by the units supported. A particular effort is made to stock parts and material for which there is the greatest demand, although quantities of items carried normally will not exceed requirements for 60 days.

Shop stores are operated in somewhat the same manner as ready issue stores except that they are located primarily in industrial or repair activity shop areas, such as in an Overhaul and Repair Department of an air station. Therefore, stock is specializd, stock control and issue are simplified, and the store is operated by personnel who are connected with the shop concerned and familiar with its material requirements.

SOURCE CODES.—When ordering a part of an electronic equipment, always check the source code listed in the parts breakdown for this may prevent much delay. The reason for this is that the item may be available only to an overhaul and repair activity. There is a special designation, or an "X" item, which means that the item is not stocked and requires the procurement of the next larger assembly, coded (P1). A description of these source codes will be found in the front of the *Illustrated Parts Breakdowns*, or the *Digest—U.S. Navy Aviation Electronics*, November, 1954, may be referred to for an explanation of the source codes.

RECORD OF MATERIAL ON ORDER.—A record of the material that you have on order will serve many functions. There will be many instances when there is confusion, loss of time, and misunderstanding in relation to whether or not a certain item has been ordered. Unless a record is kept there will be uncertainty until the item is received or correspondence regarding the order presents some evidence. Records of material ordered will not only enable you to ascertain if it was ordered, but also the priority, date, by whom ordered, and for what unit of equipment. Also, this information will aid in the installation or use of the material, for in many cases the time elapsed may be great and the man doing the work may not be present. The record that has been kept may be used as a reference. A suggested means of keeping such a record is to enter pertinent information on the back of the stub requisition that is retained in the electronics shop.

SCREENING ELECTRONIC EQUIPMENT

As a first class or chief, it will be your responsibility to determine if an electronic component should be repaired or turned into supply for credit or disposition. By utilizing spare equipments, spare parts, tools, and test equipment provided or available, using or servicing activities should exert maximum effort to effect repairs to defective equipment within their own organization.

When it becomes necessary to return material, there are prescribed procedures to be followed. The procedures are somewhat involved and require a great deal of effort but are set forth by the Supply Department of the Navy to facilitate this phase of electronics maintenance. All material that is returned should be identified with color-coded tags, for this allows a quick check as to the reason for return.

The tags with their meaning and the identification data are as follows:

1. White—Ready Issue—NavAer Form 2650 (5–51)
2. Blue—Screening—NavAer Form 2650–B
3. Green—Repairable—NavAer Form 2650–C
4. Yellow—Salvage—NavAer Form 2650–D

Electronic equipment in need of repair or overhaul is returned to Supply as prospective condition code B. This identifies the material or equipment that requires screening and/or overhaul prior to being placed in a Ready for Issue (RFI) condition.

Materials returned as excess equipment, that are in proper operation order, will be identified as RFI (Ready for Issue).

Whenever activities are unable to effect repairs to a system, equipment, or component which necessitates return to supply for disposition and the requisitioning of replacement because of the lack of parts, tools, test equipment, or inadequate facilities, a report to that effect shall be made to the Bureau of Aeronautics for analysis and study. This report may be submitted as a comment under "Remarks" on the current Electronics Failure Report by official letter or other means. The information submitted will be used for purposes of planning, correcting deficiencies, and improving field maintenance.

In the event a component or test feature of a specific item of test equipment is defective, and the cognizant activity desires to utilize temporarily the other components of the unit, the complete test equipment may be retained and a requisition submitted to the cognizant supply officer. The requisition shall bear a notation that the unit is being retained pending receipt of a replacement. For more detailed information about repair of test equipment refer to BuAer Instruction 10550.5.

ELECTRONICS FAILURE REPORT

The Electronics Failure Reports are designed to provide the Bureau of Aeronautics with up-to-date information on aeronautical electronic and electric material failures or faults and to serve as a basic tool for conducting reliability studies. Data from the reports will be transcribed by the Bureau to punched cards to permit automatic accounting machine processing, tabulation, and analysis of failure data. The analyses will be utilized as a medium for improving reliability, performance, design, and maintenance of equipment, and materials, and for contractor's study of failures and remedy of unsatisfactory conditions. Compiled data will also serve as a guide for operational and supply activities to adjust maintenance support stock levels. In order for the failure reports to serve the above purposes, it is imperative that each and every failure or fault be reported, and that reports be complete in all details.

You are encouraged to submit any suggestion that might be of value in connection with failures. Information or data which the reporting activity considers may assist in evaluating the cause of failure, or which may offer relief from additional failures, is desired. Photographs, drawings, sketches, samples, etc., may be submitted as enclosures to the basic report.

USAGE DATA

Usage data give percentages of replacement relative to hours of equipment operation and to number of overhauls. These percentages of replacement are applied to operational plans to determine probable future usage. Usage data are used in connection with the PURS (PROGRAM USAGE REPLENISHMENT SYSTEM) Program and as such is the important part of that program for it furnishes data from the field and fleet. Usage data determine the quantities of materials that will be placed on allowance lists; and with that in mind, you can easily see the importance of the information that they contain.

An Operational Data Sheet is provided for the use of activities reporting usage data. Operating maintenance activities designated by the cognizant command shall use this form as a usage and consumption report. Only ultimate consumer activities such as carriers, tenders, FASRons, self-supporting aircraft squadrons, battleships, cruisers, Reserve Air Stations, FAETU's, Air Reserve Training Units, Marine Air FMF HEDRons, Marine Service Squadrons, and Ready-Issue Stores providing direct operational support for standard organized units will be designated. All reporting activities shall fill in the appropriate spaces on the Operational Data Sheet and Column 18 throughout the list. Column 18 will reflect the total quantity of each item used from whichever source obtained in direct support of the number of assigned equipments, as shown on the Operational Data Sheet for the period covered. The number of installed equipments shall be determined by adding the number on hand each day during the reporting period and dividing the total by the number of days covered by the report.

To conserve forms, activities with low usage are encouraged to use facsimiles in lieu of the printed form. If this is done, both usage and operational data must be reported, items must be identified by item numbers from the allowance list, and the report must in all respects be the same as if the printed form were used.

All reports shall be submitted quarterly in accordance with schedules submitted by cognizant command and shall cover the preceding 90-day period or any part of that period but not less than 30 days during which normal operations were conducted. If, for any reason, a report is not submitted, the Aviation Supply Officer shall be notified by letter of explanation. It is important that care be exercised in entering correct and complete information. Completed report forms shall be submitted direct to the ASO, with a copy of the transmittal letter to the cognizant command.

Usage data are important and entail a great deal of work. An accounting of materials used must be made, so learn what should be done and how to do it. The big problem is not in filling out the necessary forms, but it is one of getting started soon enough in order to complete necessary work by the submitting date. Do not put this off until the last minute.

STANDARD AIRCRAFT INVENTORY LOG

These logs are used to facilitate the transfer of naval aircraft between activities giving assurance that all items of the aircraft and engines are intact. Inventory Logs are prepared and distributed by the Aviation Supply Office for each model plane. A logbook will be kept for each plane.

The logbook is constructed to allow for entries in certain columns. A separate column is used for each transfer and receipt of the plane. Entries are made opposite the listed equipment and show whether or not the equipment is in the plane.

SURVEY OF MATERIAL

When accountable material is damaged, worn out, missing, stolen, or otherwise not available for use, a SURVEY must be made to obtain proper authority to write this material off the books. Think of a survey as being an administrative examination into the cause of material being lost to use.

A survey may be either INFORMAL or FORMAL, depending on the circumstances in the case in question.

When a survey becomes necessary in your department, the electronics officer makes up a rough REQUEST FOR SURVEY and gives it to the supply officer. (Or it may happen that you, as a Chief Aviation Electronics Technician, will make up the rough for the electronics officer to approve.) The request lists the nature of the material and the conditions making the survey necessary. The request is made on a Survey Request, Report, and Expenditure Form (SandA Form 154). The supply officer prepares a smooth Form 154 from this rough draft and sends all the necessary papers to the commanding officer for appropriate action.

When the commanding officer receives this information, he decides which type of survey is required.

For routine matters where it appears that no disciplinary action is required, or where there are no peculiar circumstances that would require a formal review, the commanding officer will direct that an INFORMAL SURVEY be held. He will designate as the surveying officer the head of the department having custody (physical possession) of the material to be surveyed.

The commanding officer will direct that a formal survey be held by a commissioned officer or by a board of three officers, if:

1. If appears that responsibility for the loss or damage to the material may be placed on a person in the naval service.
2. The material is short or damaged in shipment by a commercial or government carrier.
3. The value of the material is in excess of $100.

QUIZ

1. Which of the following is not a function of ASO?
    a. Procurement of most aeronautical supplies
    b. Disposal of surplus materials
    c. Procurement of all aeronautical supplies
    d. General control and rationing of most aeronautical materials

2. In order to find the Stock No. of a capacitor or resistor common to three or more types of electronic equipments, the most logical place to look would be
    a. NavAer 00–35QR–7
    b. NavAer 00–35QR–4
    c. NavAer 00–35QR–6
    d. NavAer 00–35QR–5

3. Headsets, microphones, and comparable equipment would be found in
    a. NavAer 00–35QR–7
    b. NavAer 00–35QR–4
    c. NavAer 00–35QR–6
    d. NavAer 00–35QR–5

4. Stock No. for vacuum tubes are listed in
    a. NavAer 00–35QR–7
    b. NavAer 00–35QR–4
    c. NavAer 00–35QR–6
    d. NavAer 00–35QR–5

5. Source codes are used to determine which of the following?
    a. If ESO has control of the item
    b. If the item is available to your activity
    c. If ASO has control of the item
    d. If the item must be procured through open purchase

6. Materials returned as excess equipments, that are in proper operating order, will be identified
    a. RMCM
    b. RFI
    c. RUDM
    d. 00–35QR

7. All screened electronic equipment returned ready for issue will be tagged
   a. white NavAer Form 2650 (5–51)
   b. blue NavAer Form 2650–B
   c. green NavAer Form 2650–C
   d. yellow NavAer Form 2650–D

8. All screened electronic equipment returned for repairs will be tagged
   a. white NavAer Form 2650 (5–51)
   b. blue NavAer Form 2650–B
   c. green NavAer Form 2650–C
   d. yellow NavAer Form 2650–D

9. Electronic equipment in need of repair or overhaul is returned to Supply as prospective condition code
   a. A
   b. D
   c. C
   d. B

10. Which of the following are directed to turn in usage data reports?
    a. Only supply activities
    b. All using activities
    c. Only FASRon's
    d. Only designated using activities

11. Standard Aircraft Inventory Logs are prepared and distributed by
    a. ASO
    b. CNO
    c. BuSandA
    d. the aircraft manufacturer

12. A formal survey must be held when the value of the material under question exceeds
    a. $50
    b. $10
    c. $75
    d. $100

APPENDIX I

STANDARD ANNEALED SOLID COPPER WIRE TABLE

(American Wire Gage—B. and S.)

Gage number	Diameter (mils)	Cross section		Ohms per 1,000 ft.	Pounds per 1,000 ft.	Continuous current capacity (amps.)*	
		Circular mils	Square inches	25° C. (= 77° F.)		Single wire free air	Wires in conduit or bundles
0000	460.0	212,000.0	0.166	0.0500	641.0	380	225
000	410.0	168,000.0	.132	.0630	508.0	328	200
00	365.0	133,000.0	.105	.0795	403.0	283	175
0	325.0	106,000.0	.0829	.100	319.0	245	150
1	289.0	83,700.0	.0657	.126	253.0	211	125
2	258.0	66,400.0	.0521	.159	201.0	181	100
3	229.0	52,600.0	.0413	.201	159.0	-------	------
4	204.0	41,700.0	.0328	.253	126.0	135	80
5	182.0	33,100.0	.0260	.319	100.0	-------	------
6	162.0	26,300.0	.0206	.403	79.5	101	60
7	144.0	20,800.0	.0164	.508	63.0	-------	------
8	128.0	16,500.0	.0130	.641	50.0	73	46
9	114.0	13,100.0	.0103	.808	39.6	-------	------
10	102.0	10,400.0	.00815	1.02	31.4	55	33
11	91.0	8,230.0	.00647	1.28	24.9	-------	------
12	81.0	6,530.0	.00513	1.62	19.8	41	23
13	72.0	5,180.0	.00407	2.04	15.7	-------	------
14	64.0	4,110.0	.00323	2.58	12.4	32	17
15	57.0	3,260.0	.00256	3.25	9.86	-------	------
16	51.0	2,580.0	.00203	4.09	7.82	22	13
17	45.0	2,050.0	.00161	5.16	6.20	-------	------
18	40.0	1,620.0	.00128	6.51	4.92	16	10
19	36.0	1,290.0	.00101	8.21	3.90	-------	------
20	32.0	1,020.0	.000802	10.4	3.09	11	7.5
21	28.5	810.0	.000636	13.1	2.45	-------	------
22	25.3	642.0	.000505	16.5	1.94	-------	---- 5
23	22.6	509.0	.000400	20.8	1.54	-------	------
24	20.1	404.0	.000317	26.2	1.22	-------	------
25	17.9	320.0	.000252	33.0	0.970	-------	------

Appendix I continued

Gage number	Diameter (mils)	Cross section		Ohms per 1,000 ft.	Pounds per 1,000 ft.	Continuous current capacity (amps.)*	
		Circular mils	Square inches	25° C. (= 77° F.)		Single wire free air	Wires in conduit or bundles
26	15.9	254.0	.000200	41.6	0.769	-------	------
27	14.2	202.0	.000158	52.5	0.610	-------	------
28	12.6	160.0	.000126	66.2	0.484	-------	------
29	11.3	127.0	.0000995	83.4	0.384	-------	------
30	10.0	101.0	.0000789	105.0	0.304	-------	------
31	8.9	79.7	.0000626	133.0	0.241	-------	------
32	8.0	63.2	.0000496	167.0	0.191	-------	------
33	7.1	50.1	.0000394	211.0	0.152	-------	------
34	6.3	39.8	.0000312	266.0	0.120	-------	------
35	5.6	31.5	.0000248	335.0	0.0954	-------	------
36	5.0	25.0	.0000196	423.0	0.0757	-------	------
37	4.5	19.8	.0000156	533.0	0.0600	-------	------
38	4.0	15.7	.0000123	673.0	0.0476	-------	------
39	3.5	12.5	.0000098	848.0	0.0377	-------	------
40	3.1	9.9	.0000078	1,070.0	0.0299	-------	------

*Military Specification MIL–W–5088A

NOTE: The current carrying capacity of ALUMINUM wire is approximately 80% that of copper.

APPENDIX II

GREEK ALPHABET

Name	Capital	Lower case	Designates
Alpha	Α	α	Angles.
Beta	Β	β	Angles, flux density.
Gamma	Γ	γ	Conductivity.
Delta	Δ	δ	Variation of a quantity, increment.
Epsilon	Ε	ε	Base of natural logarithms (2.71828).
Zeta	Ζ	ζ	Impedance, coefficients, coordinates.
Eta	Η	η	Hysteresis coefficient, efficiency.
Theta	Θ	θ	Phase angle.
Iota	Ι	ι	
Kappa	Κ	κ	Dielectric constant, coupling coefficient, susceptibility.
Lambda	Λ	λ	Wavelength.
Mu	Μ	μ	Permeability, micro, amplification factor.
Nu	Ν	ν	Reluctivity.
Xi	Ξ	ξ	
Omicon	Ο	ο	
Pi	Π	π	3.1416
Rho	Ρ	ρ	Resistivity.
Sigma	Σ	σ	
Tau	Τ	τ	Time constant, time-phase displacement.
Upsilon	Υ	υ	
Phi	Φ	φ	Angles, magnetic flux.
Chi	Χ	χ	
Psi	Ψ	ψ	Dielectric flux, phase difference.
Omega	Ω	ω	Ohms (capital), angular velocity (2 π f).

APPENDIX III

FORMULAS

Ohm's Law for D-C Circuits

$$I=\frac{E}{R}=\frac{P}{E}=\sqrt{\frac{P}{R}}$$

$$R=\frac{E}{I}=\frac{P}{I^2}=\frac{E^2}{P}$$

$$E=IR=\frac{P}{I}=\sqrt{PR}$$

$$P=EI=\frac{E^2}{R}=I^2R$$

Resistors in Series

$$R_T=R_1+R_2 \ldots$$

Resistors in Parallel

Two resistors

$$R_T=\frac{R_1R_2}{R_1+R_2}$$

More than two

$$\frac{1}{R_T}=\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3} \ldots$$

R–L Circuit Time Constant equals

$$\frac{L \text{ (in henrys)}}{R \text{ (in ohms)}}=t \text{ (in seconds), or}$$

$$\frac{L \text{ (in microhenrys)}}{R \text{ (in ohms)}}=t \text{ (in microseconds)}$$

R–C Circuit Time Constants equals

R (ohms) $\times C$ (farads) $=t$ (seconds)

R (megohms) $\times C$ (microfarads) $=t$ (seconds)

R (ohms) $\times C$ (microfarads) $=t$ (microseconds)

R (megohms) $\times C$ (micromicrofarads) $=t$ (microseconds)

Comparison of Units in Electric and Magnetic Circuits

	Electric circuit	Magnetic circuit
Force	Volt, E, or emf	Gilberts, F, or mmf
Flow	Ampere, I	Flux, Φ, in maxwells
Opposition	Ohms, R	Reluctance, R
Law	Ohm's law, $I=\frac{E}{R}$	Rowland's law, $\Phi=\frac{F}{R}$
Intensity of force.	Volts per cm of length	$H=\frac{1.257IN}{L}$, gilberts per centimeter of length
Density	Current density—for example, amperes per cm^2.	Flux density—for example, lines per cm^2, or gausses

Capacitors in Series

Two capacitors

$$C_T=\frac{C_1C_2}{C_1+C_2}$$

More than two

$$\frac{1}{C_T}=\frac{1}{C_1}+\frac{1}{C_2}+\frac{1}{C_3}\ldots$$

Capacitors in Parallel

$$C_T=C_1+C_2\ldots$$

Capacitive Reactance

$$X_C=\frac{1}{2\pi fC}$$

Impedance in an R–C Circuit (series)

$$Z=\sqrt{R^2+X_C^2}$$

Inductors in Series

$L_T=L_1+L_2\ldots$ (No coupling between coils)

Inductors in Parallel

Two inductors

$$L_T = \frac{L_1 L_2}{L_1 + L_2} \text{ (No coupling between coils)}$$

More than two

$$\frac{1}{L_T} = \frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3} \dots \text{ (No coupling between coils)}$$

Inductive Reactance

$$X_L = 2\pi f L$$

Q of a Coil

$$Q = \frac{X_L}{R}$$

Impedance of an R–L Circuit (series)

$$Z = \sqrt{R^2 + X_L^2}$$

Impedance with R, C, and L in Series

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$

Parallel Circuit Impedance

$$Z = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

Sine-Wave Voltage Relationships

Average value

$$E_{\text{ave}} = \frac{2}{\pi} \times E_{\text{max}} = 0.637 E_{\text{max}}$$

Effective or r.m.s. value

$$E_{\text{eff}} = \frac{E_{\text{max}}}{\sqrt{2}} = \frac{E_{\text{max}}}{1.414} = 0.707 E_{\text{max}} = 1.11 E_{\text{ave}}$$

Maximum value

$$E_{\text{max}} = \sqrt{2}\ E_{\text{eff}} = 1.414 E_{\text{eff}} = 1.57 E_{\text{ave}}$$

Voltage in an a-c circuit

$$E = IZ = \frac{P}{I \times P.F.}$$

Current in an a-c circuit

$$I=\frac{E}{Z}=\frac{P}{E\times P.F.}$$

Power in A-C Circuit

Apparent power$=EI$

True power

$$P=EI\cos\theta=EI\times P.F.$$

Power factor

$$P.F.=\frac{P}{EI}=\cos\theta$$

$$\cos\theta=\frac{\text{true power}}{\text{apparent power}}$$

Transformers

Voltage relationship

$$\frac{E_p}{E_s}=\frac{N_p}{N_s}\text{ or }E_s=E_p\times\frac{N_s}{N_p}$$

Current relationship

$$\frac{I_p}{I_s}=\frac{N_s}{N_p}$$

Induced voltage

$$E_{eff}=4.44\ BAfN10^{-8}$$

Turns ratio equals

$$\frac{N_p}{N_s}=\sqrt{\frac{Z_p}{Z_s}}$$

Secondary current

$$I_s=I_p\frac{N_p}{N_s}$$

Secondary voltage

$$E_s=E_p\frac{N_s}{N_p}$$

Three Phase Voltage and Current Relationships

With wye connected windings

$$E_{\text{line}} = 1.732E_{\text{coil}} = \sqrt{3}\ E_{\text{coil}}$$

$$I_{\text{line}} = I_{\text{coil}}$$

With delta connected windings

$$E_{\text{line}} = E_{\text{coil}}$$

$$I_{\text{line}} = 1.732I_{\text{coil}}$$

With wye or delta connected winding

$$P_{\text{coil}} = E_{\text{coil}}I_{\text{coil}}$$

$$P_t = 3P_{\text{coil}}$$

$$P_t = 1.732E_{\text{line}}I_{\text{line}}$$

(To convert to true power multiply by cos θ)

Resonance

At resonance

$$X_L = X_C$$

Resonant frequency

$$F_o = \frac{1}{2\pi\sqrt{LC}}$$

Series resonance

$$Z \text{ (at any frequency)} = R + j(X_L - X_C)$$

$$Z \text{ (at resonance)} = R$$

Parallel resonance

$$Z \text{ (at any frequency)} = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

$$Z_{\max} \text{ (at resonance)} = \frac{X_L X_C}{R} = \frac{X_L^2}{R} = QX_L = \frac{L}{CR}$$

Band width

$$\Delta = \frac{\text{F}_o}{Q} = \frac{R}{2\pi L}$$

Tube Characteristics

Amplification factor

$$\mu = \frac{\Delta e_p}{\Delta e_g} \; (i_p \text{ constant})$$

$$\mu = g_m r_p$$

A-c plate resistance

$$r_p = \frac{\Delta e_p}{\Delta i_p} \; (e_g \text{ constant})$$

Grid-plate transconductance

$$g_m = \frac{\Delta i_p}{\Delta e_g} \; (e_p \text{ constant}).$$

Decibels

Power ratio

$$\text{db} = 10 \log_{10} \frac{P_2}{P_1}$$

Current and voltage ratio

$$\text{db} = 20 \log_{10} \frac{I_2 \sqrt{R_2}}{I_1 \sqrt{R_1}}$$

$$\text{db} = 20 \log_{10} \frac{E_2 \sqrt{R_1}}{E_1 \sqrt{R_2}}$$

NOTE: When R_1 and R_2 are equal they may be omitted fron the formula.

When reference level is one milliwatt

$$\text{dbm} = 10 \log_{10} \frac{P}{0.001} \; (\text{when } P \text{ is in watts})$$

Synchronous Speed of Motor

$$\text{r.p.m.} = \frac{120 \times \text{frequency}}{\text{number of poles}}$$

Radar Power Relationships

$$\text{Duty cycle} = \frac{\text{average power}}{\text{peak power}} = \frac{\text{pulse width}}{\text{pulse repetition time}}$$

$$\text{Average power} = \text{peak power} \times \text{duty cycle}$$

$$\text{Peak power} = \frac{\text{average power} \times 10^{6}}{\text{pulse width in } \mu \text{ sec.} \times \text{PRF}}$$

APPENDIX IV

LAWS OF EXPONENTS

The following table shows some multiples of 10 and their equivalent powers of ten which are used to write a number in scientific form.

Number	*Powers of Ten*	*Number*	*Powers of Ten*
100,000=	10^5	0.00001=	10^{-5}
10,000=	10^4	0.0001 =	10^{-4}
1,000=	10^3	0.001 =	10^{-3}
100=	10^2	0.01 =	10^{-2}
10=	10^1	0.1 =	10^{-1}
1=	10^0	1 =	10^0

To multiply like (with same base) exponential quantities, add the exponents. In the language of algebra the rule is $a^m \times a^n = a^{m+n}$.

$$10^4 \times 10^2 = 10^{4+2} = 10^6$$

$$0.003 \times 825.2 = 3 \times 10^{-3} \times 8.252 \times 10^2 = 24.756 \times 10^{-1} = 2.4756$$

To divide like exponential quantities, subtract the exponents. In the language of algebra the rule is $\frac{a^m}{a^n} = a^{m-n}$

$$10^8 \div 10^2 = 10^6$$

$$3{,}000 \div 0.015 = (3 \times 10^3) \div (1.5 \times 10^{-2}) = 2 \times 10^5 = 200{,}000$$

To raise an exponential quantity to a power, multiply the exponents. In the language of algebra $(x^m)^n = x^{mn}$.

$$(10^3)^4 = 10^{3 \times 4} = 10^{12}$$

$$2{,}500^2 = (2.5 \times 10^3)^2 = 6.25 \times 10^6 = 6{,}250{,}000$$

Any number (except zero) raised to the zero power is one. In the language of algebra $x^0 = 1$

$$x^3 \div x^3 = 1$$

$$10^4 \div 10^4 = 1$$

Any base with a negative exponent is equal to 1 divided by the base with an equal positive exponent. In the language of algebra

$$x^{-a}=\frac{1}{x^a}$$

$$10^{-2}=\frac{1}{10^2}=\frac{1}{100}$$

$$5a^{-3}=\frac{5}{a^3}$$

$$(6a)^{-1}=\frac{1}{6a}$$

To raise a product to a power, raise each factor of the product to that power.

$$(2\times10)^2=2^2\times10^2$$

$$3{,}000^3=(3\times10^3)^3=27\times10^9$$

To find the nth root of an exponential quantity, divide the exponent by the index of the root. Thus, the nth root of $a^m=a^{m/n}$.

$$\sqrt{x^6}=x^{6/2}=x^3$$

$$\sqrt[3]{64\times10^3}=4\times10=40$$

APPENDIX V

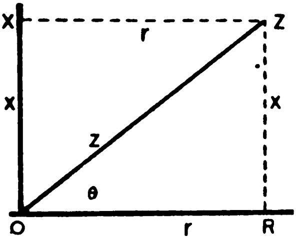
TRIGONOMETRIC FUNCTIONS

Several special relationships, called trigonometric functions, hold true in a right triangle. Electrical problems when reduced to a right triangle can be easily and quickly solved by use of tables based upon these functions.

In the diagram shown, θ is the angle ZOR, OR is the projection of OZ on the horizontal axis, OX is the projection of OZ on the vertical axis. The letters r, x, and z represent the lengths of OR, OX, and OZ respectively. There are, in all, six different ratios between the sides r, x, and z. They are called trigonometric ratios or trigonometric functions.

Sine of θ (written sin θ)$=\frac{x}{z}$

Cosine of θ (written cos θ)$=\frac{r}{z}$



Tangent of θ (written tan θ)$=\frac{x}{r}$

Cotangent of θ (written cot θ)$=\frac{r}{x}$

Secant of θ (written sec θ)$=\frac{z}{r}$

Cosecant of θ (written csc θ)$=\frac{z}{x}$

The functions, sin θ, cos θ, and sec θ, are the reciprocals of cot θ, sec θ, and csc θ, respectively. Only the first three, sin θ, cos θ, and tan θ, are generally used.

Suppose that in the diagram, OZ has a unit length of 1 and is rotated in a counterclockwise direction beginning with angle θ at 0° value and continuing until it is 90°, then the functions will vary within the following limits:

sin θ increases from 0 to 1.0

cos θ decreases from 1.0 to 0

tan θ increases from 0 to ∞

$\cot\theta$ decreases from ∞ to 0

$\sec\theta$ increases from 1.0 to ∞

$\csc\theta$ decreases from ∞ to 1.0

When θ varies between the values of 90° and 180° (second quadrant), the projection r is negative and the functions which involve r become negative. Thus, $\cos\theta$ and $\tan\theta$ are negative in this quadrant. In the third quadrant, both r and x are negative. Therefore, $\sin\theta$ and $\cos\theta$ which involve only one or the other, are negative, while $\tan\theta$, which involves both r and x, is positive. In the fourth quadrant, r is positive, but x is still negative. For this reason, $\sin\theta$ and $\tan\theta$ are negative in the fourth quadrant, while $\cos\theta$ is positive.

Variations in Value of Functions

Quadrant	Sin θ		Cos θ		Tan θ	
	From	To	From	To	From	To
I (0° –90°)	0	1.0	1.0	0	0	∞
II (90° –180°)	1.0	0	0	–1.0	–∞	0
III (180° –270°)	0	–1.0	–1.0	0	0	∞
IV (270° –360°)	–1.0	0	0	1.0	–∞	0

Trigonometric tables give functions up to 90° only. The following rules apply for functions of angles greater than 90°.

In quadrant II: $\theta = 90° +$ some angle which is designated α (alpha).

$$\sin(90°+\alpha) = \sin(90°-\alpha)$$

$$\cos(90°+\alpha) = -\cos(90°-\alpha)$$

$$\tan(90°+\alpha) = -\tan(90°-\alpha)$$

In quadrant III: $\theta = 180° + \alpha$

$$\sin(180°+\alpha) = -\sin\alpha$$

$$\cos(180°+\alpha) = -\cos\alpha$$

$$\tan(180°+\alpha) = \tan\alpha$$

In quadrant IV: $\theta = 270° + \alpha$

$$\sin(270° + \alpha) = -\sin(90° - \alpha)$$

$$\cos(270° + \alpha) = \cos(90° - \alpha)$$

$$\tan(270° + \alpha) = -\tan(90° - \alpha)$$

APPENDIX VI

Natural Sines, Cosines, and Tangents

0°–14.9°

Degs.	Function	0.0°	0.1°	0.2°	0.3°	0.4°	0.5°	0.6°	0.7°	0.8°	0.9°
0	sin	0.0000	0.0017	0.0035	0.0052	0.0070	0.0087	0.0105	0.0122	0.0140	0.0157
	cos	1.0000	1.0000	1.0000	1.0000	1.0000	1.0000	0.9999	0.9999	0.9999	0.9999
	tan	0.0000	0.0017	0.0035	0.0052	0.0070	0.0087	0.0105	0.0122	0.0140	0.0157
1	sin	0.0175	0.0192	0.0209	0.0227	0.0244	0.0262	0.0279	0.0297	0.0314	0.0332
	cos	0.9998	0.9998	0.9998	0.9997	0.9997	0.9997	0.9996	0.9996	0.9995	0.9995
	tan	0.0175	0.0192	0.0209	0.0227	0.0244	0.0262	0.0279	0.0297	0.0314	0.0332
2	sin	0.0349	0.0366	0.0384	0.0401	0.0419	0.0436	0.0454	0.0471	0.0488	0.0506
	cos	0.9994	0.9993	0.9993	0.9992	0.9991	0.9990	0.9990	0.9989	0.9988	0.9987
	tan	0.0349	0.0367	0.0384	0.0402	0.0419	0.0437	0.0454	0.0472	0.0489	0.0507
3	sin	0.0523	0.0541	0.0558	0.0576	0.0593	0.0610	0.0628	0.0645	0.0663	0.0680
	cos	0.9986	0.9985	0.9984	0.9983	0.9982	0.9981	0.9980	0.9979	0.9978	0.9977
	tan	0.0524	0.0542	0.0559	0.0577	0.0594	0.0612	0.0629	0.0647	0.0664	0.0682
4	sin	0.0698	0.0715	0.0732	0.0750	0.0767	0.0785	0.0802	0.0819	0.0837	0.0854
	cos	0.9976	0.9974	0.9973	0.9972	0.9971	0.9969	0.9968	0.9966	0.9965	0.9963
	tan	0.0699	0.0717	0.0734	0.0752	0.0769	0.0787	0.0805	0.0822	0.0840	0.0857
5	sin	0.0872	0.0889	0.0906	0.0924	0.0941	0.0958	0.0976	0.0993	0.1011	0.1028
	cos	0.9962	0.9960	0.9959	0.9957	0.9956	0.9954	0.9952	0.9951	0.9949	0.9947
	tan	0.0875	0.0892	0.0910	0.0928	0.0945	0.0963	0.0981	0.0998	0.1016	0.1033
6	sin	0.1045	0.1063	0.1080	0.1097	0.1115	0.1132	0.1149	0.1167	0.1184	0.1201
	cos	0.9945	0.9943	0.9942	0.9940	0.9938	0.9936	0.9934	0.9932	0.9930	0.9928
	tan	0.1051	0.1069	0.1086	0.1104	0.1122	0.1139	0.1157	0.1175	0.1192	0.1210
7	sin	0.1219	0.1236	0.1253	0.1271	0.1288	0.1305	0.1323	0.1340	0.1357	0.1374
	cos	0.9925	0.9923	0.9921	0.9919	0.9917	0.9914	0.9912	0.9910	0.9907	0.9905
	tan	0.1228	0.1246	0.1263	0.1281	0.1299	0.1317	0.1334	0.1352	0.1370	0.1388
8	sin	0.1392	0.1409	0.1426	0.1444	0.1461	0.1478	0.1495	0.1513	0.1530	0.1547
	cos	0.9903	0.9900	0.9898	0.9895	0.9893	0.9890	0.9888	0.9885	0.9882	0.9880
	tan	0.1405	0.1423	0.1441	0.1459	0.1477	0.1495	0.1512	0.1530	0.1548	0.1566
9	sin	0.1564	0.1582	0.1599	0.1616	0.1633	0.1650	0.1668	0.1685	0.1702	0.1719
	cos	0.9877	0.9874	0.9871	0.9869	0.9866	0.9863	0.9860	0.9857	0.9854	0.9851
	tan	0.1584	0.1602	0.1620	0.1638	0.1655	0.1673	0.1691	0.1709	0.1727	0.1745
10	sin	0.1736	0.1754	0.1771	0.1788	0.1805	0.1822	0.1840	0.1857	0.1874	0.1891
	cos	0.9848	0.9845	0.9842	0.9839	0.9836	0.9833	0.9829	0.9826	0.9823	0.9820
	tan	0.1763	0.1781	0.1799	0.1817	0.1835	0.1853	0.1871	0.1890	0.1908	0.1926
11	sin	0.1908	0.1925	0.1942	0.1959	0.1977	0.1994	0.2011	0.2028	0.2045	0.2062
	cos	0.9816	0.9813	0.9810	0.9806	0.9803	0.9799	0.9796	0.9792	0.9789	0.9785
	tan	0.1944	0.1962	0.1980	0.1998	0.2016	0.2035	0.2053	0.2071	0.2089	0.2107
12	sin	0.2079	0.2096	0.2113	0.2130	0.2147	0.2164	0.2181	0.2198	0.2215	0.2232
	cos	0.9781	0.9778	0.9774	0.9770	0.9767	0.9763	0.9759	0.9755	0.9751	0.9748
	tan	0.2126	0.2144	0.2162	0.2180	0.2199	0.2217	0.2235	0.2254	0.2272	0.2290
13	sin	0.2250	0.2267	0.2284	0.2300	0.2318	0.2334	0.2351	0.2368	0.2385	0.2402
	cos	0.9744	0.9740	0.9736	0.9732	0.9728	0.9724	0.9720	0.9715	0.9711	0.9707
	tan	0.2309	0.2327	0.2345	0.2364	0.2382	0.2401	0.2419	0.2438	0.2456	0.2475
14	sin	0.2419	0.2436	0.2453	0.2470	0.2487	0.2504	0.2521	0.2538	0.2554	0.2571
	cos	0.9703	0.9699	0.9694	0.9690	0.9686	0.9681	0.9677	0.9673	0.9668	0.9664
	tan	0.2493	0.2512	0.2530	0.2549	0.2568	0.2586	0.2605	0.2623	0.2642	0.2661
Degs.	Function	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′

Degs.	Function	0.0°	0.1°	0.2°	0.3°	0.4°	0.5°	0.6°	0.7°	0.8°	0.9°
15	sin	0.2588	0.2605	0.2622	0.2639	0.2656	0.2672	0.2689	0.2706	0.2723	0.2740
	cos	0.9659	0.9655	0.9650	0.9646	0.9641	0.9636	0.9632	0.9627	0.9622	0.9617
	tan	0.2679	0.2698	0.2717	0.2736	0.2754	0.2773	0.2792	0.2811	0.2830	0.2849
16	sin	0.2756	0.2773	0.2790	0.2807	0.2823	0.2840	0.2857	0.2874	0.2890	0.2907
	cos	0.9613	0.9608	0.9603	0.9598	0.9593	0.9588	0.9583	0.9578	0.9573	0.9568
	tan	0.2867	0.2886	0.2905	0.2924	0.2943	0.2962	0.2981	0.3000	0.3019	0.3038
17	sin	0.2924	0.2940	0.2957	0.2974	0.2990	0.3007	0.3024	0.3040	0.3057	0.3074
	cos	0.9563	0.9558	0.9553	0.9548	0.9542	0.9537	0.9532	0.9527	0.9521	0.9516
	tan	0.3057	0.3076	0.3096	0.3115	0.3134	0.3153	0.3172	0.3191	0.3211	0.3230
18	sin	0.3090	0.3107	0.3123	0.3140	0.3156	0.3173	0.3190	0.3206	0.3223	0.3239
	cos	0.9511	0.9505	0.9500	0.9494	0.9489	0.9483	0.9478	0.9472	0.9466	0.9461
	tan	0.3249	0.3269	0.3288	0.3307	0.3327	0.3346	0.3365	0.3385	0.3404	0.3424
19	sin	0.3256	0.3272	0.3289	0.3305	0.3322	0.3338	0.3355	0.3371	0.3387	0.3404
	cos	0.9455	0.9449	0.9444	0.9438	0.9432	0.9426	0.9421	0.9415	0.9409	0.9403
	tan	0.3443	0.3463	0.3482	0.3502	0.3522	0.3541	0.3561	0.3581	0.3600	0.3620
20	sin	0.3420	0.3437	0.3453	0.3469	0.3486	0.3502	0.3518	0.3535	0.3551	0.3567
	cos	0.9397	0.9391	0.9385	0.9379	0.9373	0.9367	0.9361	0.9354	0.9348	0.9342
	tan	0.3640	0.3659	0.3679	0.3699	0.3719	0.3739	0.3759	0.3779	0.3799	0.3819
21	sin	0.3584	0.3600	0.3616	0.3633	0.3649	0.3665	0.3681	0.3697	0.3714	0.3730
	cos	0.9336	0.9330	0.9323	0.9317	0.9311	0.9304	0.9298	0.9291	0.9285	0.9278
	tan	0.3839	0.3859	0.3879	0.3899	0.3919	0.3939	0.3959	0.3979	0.4000	0.4020
22	sin	0.3746	0.3762	0.3778	0.3795	0.3811	0.3827	0.3843	0.3859	0.3875	0.3891
	cos	0.9272	0.9265	0.9259	0.9252	0.9245	0.9239	0.9232	0.9225	0.9219	0.9212
	tan	0.4040	0.4061	0.4081	0.4101	0.4122	0.4142	0.4163	0.4183	0.4204	0.4224
23	sin	0.3907	0.3923	0.3939	0.3955	0.3971	0.3987	0.4003	0.4019	0.4035	0.4051
	cos	0.9205	0.9198	0.9191	0.9184	0.9178	0.9171	0.9164	0.9157	0.9150	0.9143
	tan	0.4245	0.4265	0.4286	0.4307	0.4327	0.4348	0.4369	0.4390	0.4411	0.4431
24	sin	0.4067	0.4083	0.4099	0.4115	0.4131	0.4147	0.4163	0.4179	0.4195	0.4210
	cos	0.9135	0.9128	0.9121	0.9114	0.9107	0.9100	0.9092	0.9085	0.9078	0.9070
	tan	0.4452	0.4473	0.4494	0.4515	0.4536	0.4557	0.4578	0.4599	0.4621	0.4642
25	sin	0.4226	0.4242	0.4258	0.4274	0.4289	0.4305	0.4321	0.4337	0.4352	0.4368
	cos	0.9063	0.9056	0.9048	0.9041	0.9033	0.9026	0.9018	0.9011	0.9003	0.8996
	tan	0.4663	0.4684	0.4706	0.4727	0.4748	0.4770	0.4791	0.4813	0.4834	0.4856
26	sin	0.4384	0.4399	0.4415	0.4431	0.4446	0.4462	0.4478	0.4493	0.4509	0.4524
	cos	0.8988	0.8980	0.8973	0.8965	0.8957	0.8949	0.8942	0.8934	0.8926	0.8918
	tan	0.4877	0.4899	0.4921	0.4942	0.4964	0.4986	0.5008	0.5029	0.5051	0.5073
27	sin	0.4540	0.4555	0.4571	0.4586	0.4602	0.4617	0.4633	0.4648	0.4664	0.4679
	cos	0.8910	0.8902	0.8894	0.8886	0.8878	0.8870	0.8862	0.8854	0.8846	0.8838
	tan	0.5095	0.5117	0.5139	0.5161	0.5184	0.5206	0.5228	0.5250	0.5272	0.5295
28	sin	0.4695	0.4710	0.4726	0.4741	0.4756	0.4772	0.4787	0.4802	0.4818	0.4833
	cos	0.8829	0.8821	0.8813	0.8805	0.8796	0.8788	0.8780	0.8771	0.8763	0.8755
	tan	0.5317	0.5340	0.5362	0.5384	0.5407	0.5430	0.5452	0.5475	0.5498	0.5520
29	sin	0.4848	0.4863	0.4879	0.4894	0.4909	0.4924	0.4939	0.4955	0.4970	0.4985
	cos	0.8746	0.8738	0.8729	0.8721	0.8712	0.8704	0.8695	0.8686	0.8678	0.8669
	tan	0.5543	0.5566	0.5589	0.5612	0.5635	0.5658	0.5681	0.5704	0.5727	0.5750
Degs.	Function	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′

NATURAL SINES, COSINES, AND TANGENTS

Degs.	Function	0.0°	0.1°	0.2°	0.3°	0.4°	0.5°	0.6°	0.7°	0.8°	0.9°
30	sin	0.5000	0.5015	0.5030	0.5045	0.5060	0.5075	0.5090	0.5105	0.5120	0.5135
	cos	0.8660	0.8652	0.8643	0.8634	0.8625	0.8616	0.8607	0.8599	0.8590	0.8581
	tan	0.5774	0.5797	0.5820	0.5844	0.5867	0.5890	0.5914	0.5938	0.5961	0.5985
31	sin	0.5150	0.5165	0.5180	0.5195	0.5210	0.5225	0.5240	0.5255	0.5270	0.5284
	cos	0.8572	0.8563	0.8554	0.8545	0.8536	0.8526	0.8517	0.8508	0.8499	0.8490
	tan	0.6009	0.6032	0.6056	0.6080	0.6104	0.6128	0.6152	0.6176	0.6200	0.6224
32	sin	0.5299	0.5314	0.5329	0.5344	0.5358	0.5373	0.5388	0.5402	0.5417	0.5432
	cos	0.8480	0.8471	0.8462	0.8453	0.8443	0.8434	0.8425	0.8415	0.8406	0.8396
	tan	0.6249	0.6273	0.6297	0.6322	0.6346	0.6371	0.6395	0.6420	0.6445	0.6469
33	sin	0.5446	0.5461	0.5476	0.5490	0.5505	0.5519	0.5534	0.5548	0.5563	0.5577
	cos	0.8387	0.8377	0.8368	0.8358	0.8348	0.8339	0.8329	0.8320	0.8310	0.8300
	tan	0.6494	0.6519	0.6544	0.6569	0.6594	0.6619	0.6644	0.6669	0.6694	0.6720
34	sin	0.5592	0.5606	0.5621	0.5635	0.5650	0.5664	0.5678	0.5693	0.5707	0.5721
	cos	0.8290	0.8281	0.8271	0.8261	0.8251	0.8241	0.8231	0.8221	0.8211	0.8202
	tan	0.6745	0.6771	0.6796	0.6822	0.6847	0.6873	0.6899	0.6924	0.6950	0.6976
35	sin	0.5736	0.5750	0.5764	0.5779	0.5793	0.5807	0.5821	0.5835	0.5850	0.5864
	cos	0.8192	0.8181	0.8171	0.8161	0.8151	0.8141	0.8131	0.8121	0.8111	0.8100
	tan	0.7002	0.7028	0.7054	0.7080	0.7107	0.7133	0.7159	0.7186	0.7212	0.7239
36	sin	0.5878	0.5892	0.5906	0.5920	0.5934	0.5948	0.5962	0.5976	0.5990	0.6004
	cos	0.8090	0.8080	0.8070	0.8059	0.8049	0.8039	0.8028	0.8018	0.8007	0.7997
	tan	0.7265	0.7292	0.7319	0.7346	0.7373	0.7400	0.7427	0.7454	0.7481	0.7508
37	sin	0.6018	0.6032	0.6046	0.6060	0.6074	0.6088	0.6101	0.6115	0.6129	0.6143
	cos	0.7986	0.7976	0.7965	0.7955	0.7944	0.7934	0.7923	0.7912	0.7902	0.7891
	tan	0.7536	0.7563	0.7590	0.7618	0.7646	0.7673	0.7701	0.7729	0.7757	0.7785
38	sin	0.6157	0.6170	0.6184	0.6198	0.6211	0.6225	0.6239	0.6252	0.6266	0.6280
	cos	0.7880	0.7869	0.7859	0.7848	0.7837	0.7826	0.7815	0.7804	0.7793	0.7782
	tan	0.7813	0.7841	0.7869	0.7898	0.7926	0.7954	0.7983	0.8012	0.8040	0.8069
39	sin	0.6293	0.6307	0.6320	0.6334	0.6347	0.6361	0.6374	0.6388	0.6401	0.6414
	cos	0.7771	0.7760	0.7749	0.7738	0.7727	0.7716	0.7705	0.7694	0.7683	0.7672
	tan	0.8098	0.8127	0.8156	0.8185	0.8214	0.8243	0.8273	0.8302	0.8332	0.8361
40	sin	0.6428	0.6441	0.6455	0.6468	0.6481	0.6494	0.6508	0.6521	0.6534	0.6547
	cos	0.7660	0.7649	0.7638	0.7627	0.7615	0.7604	0.7593	0.7581	0.7570	0.7559
	tan	0.8391	0.8421	0.8451	0.8481	0.8511	0.8541	0.8571	0.8601	0.8632	0.8662
41	sin	0.6561	0.6574	0.6587	0.6600	0.6613	0.6626	0.6639	0.6652	0.6665	0.6678
	cos	0.7547	0.7536	0.7524	0.7513	0.7501	0.7490	0.7478	0.7466	0.7455	0.7443
	tan	0.8693	0.8724	0.8754	0.8785	0.8816	0.8847	0.8878	0.8910	0.8941	0.8972
42	sin	0.6691	0.6704	0.6717	0.6730	0.6743	0.6756	0.6769	0.6782	0.6794	0.6807
	cos	0.7431	0.7420	0.7408	0.7396	0.7385	0.7373	0.7361	0.7349	0.7337	0.7325
	tan	0.9004	0.9036	0.9067	0.9099	0.9131	0.9163	0.9195	0.9228	0.9260	0.9293
43	sin	0.6820	0.6833	0.6845	0.6858	0.6871	0.6884	0.6896	0.6909	0.6921	0.6934
	cos	0.7314	0.7302	0.7290	0.7278	0.7266	0.7254	0.7242	0.7230	0.7218	0.7206
	tan	0.9325	0.9358	0.9391	0.9424	0.9457	0.9490	0.9523	0.9556	0.9590	0.9623
44	sin	0.6947	0.6959	0.6972	0.6984	0.6997	0.7009	0.7022	0.7034	0.7046	0.7059
	cos	0.7193	0.7181	0.7169	0.7157	0.7145	0.7133	0.7120	0.7108	0.7096	0.7083
	tan	0.9657	0.9691	0.9725	0.9759	0.9793	0.9827	0.9861	0.9896	0.9930	0.9965
Degs.	Function	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′

Degs.	Function	0.0°	0.1°	0.2°	0.3°	0.4°	0.5°	0.6°	0.7°	0.8°	0.9°
45	sin	0.7071	0.7083	0.7096	0.7108	0.7120	0.7133	0.7145	0.7157	0.7169	0.7181
	cos	0.7071	0.7059	0.7046	0.7034	0.7022	0.7009	0.6997	0.6984	0.6972	0.6959
	tan	1.0000	1.0035	1.0070	1.0105	1.0141	1.0176	1.0212	1.0247	1.0283	1.0319
46	sin	0.7193	0.7206	0.7218	0.7230	0.7242	0.7254	0.7266	0.7278	0.7290	0.7302
	cos	0.6947	0.6934	0.6921	0.6909	0.6896	0.6884	0.6871	0.6858	0.6845	0.6833
	tan	1.0355	1.0392	1.0428	1.0464	1.0501	1.0538	1.0575	1.0612	1.0649	1.0686
47	sin	0.7314	0.7325	0.7337	0.7349	0.7361	0.7373	0.7385	0.7396	0.7408	0.7420
	cos	0.6820	0.6807	0.6794	0.6782	0.6769	0.6756	0.6743	0.6730	0.6717	0.6704
	tan	1.0724	1.0761	1.0799	1.0837	1.0875	1.0913	1.0951	1.0990	1.1028	1.1067
48	sin	0.7431	0.7443	0.7455	0.7466	0.7478	0.7490	0.7501	0.7513	0.7524	0.7536
	cos	0.6691	0.6678	0.6665	0.6652	0.6639	0.6626	0.6613	0.6600	0.6587	0.6574
	tan	1.1106	1.1145	1.1184	1.1224	1.1263	1.1303	1.1343	1.1383	1.1423	1.1463
49	sin	0.7547	0.7559	0.7570	0.7581	0.7593	0.7604	0.7615	0.7627	0.7638	0.7649
	cos	0.6561	0.6547	0.6534	0.6521	0.6508	0.6494	0.6481	0.6468	0.6455	0.6441
	tan	1.1504	1.1544	1.1585	1.1626	1.1667	1.1708	1.1750	1.1792	1.1833	1.1875
50	sin	0.7660	0.7672	0.7683	0.7694	0.7705	0.7716	0.7727	0.7738	0.7749	0.7760
	cos	0.6428	0.6414	0.6401	0.6388	0.6374	0.6361	0.6347	0.6334	0.6320	0.6307
	tan	1.1918	1.1960	1.2002	1.2045	1.2088	1.2131	1.2174	1.2218	1.2261	1.2305
51	sin	0.7771	0.7782	0.7793	0.7804	0.7815	0.7826	0.7837	0.7848	0.7859	0.7869
	cos	0.6293	0.6280	0.6266	0.6252	0.6239	0.6225	0.6211	0.6198	0.6184	0.6170
	tan	1.2349	1.2393	1.2437	1.2482	1.2527	1.2572	1.2617	1.2662	1.2708	1.2753
52	sin	0.7880	0.7891	0.7902	0.7912	0.7923	0.7934	0.7944	0.7955	0.7965	0.7976
	cos	0.6157	0.6143	0.6129	0.6115	0.6101	0.6088	0.6074	0.6060	0.6046	0.6032
	tan	1.2799	1.2846	1.2892	1.2938	1.2985	1.3032	1.3079	1.3127	1.3175	1.3222
53	sin	0.7986	0.7997	0.8007	0.8018	0.8028	0.8039	0.8049	0.8059	0.8070	0.8080
	cos	0.6018	0.6004	0.5990	0.5976	0.5962	0.5948	0.5934	0.5920	0.5906	0.5892
	tan	1.3270	1.3319	1.3367	1.3416	1.3465	1.3514	1.3564	1.3613	1.3663	1.3713
54	sin	0.8090	0.8100	0.8111	0.8121	0.8131	0.8141	0.8151	0.8161	0.8171	0.8181
	cos	0.5878	0.5864	0.5850	0.5835	0.5821	0.5807	0.5793	0.5779	0.5764	0.5750
	tan	1.3764	1.3814	1.3865	1.3916	1.3968	1.4019	1.4071	1.4124	1.4176	1.4229
55	sin	0.8192	0.8202	0.8211	0.8221	0.8231	0.8241	0.8251	0.8261	0.8271	0.8281
	cos	0.5736	0.5721	0.5707	0.5693	0.5678	0.5664	0.5650	0.5635	0.5621	0.5606
	tan	1.4281	1.4335	1.4388	1.4442	1.4496	1.4550	1.4605	1.4659	1.4715	1.4770
56	sin	0.8290	0.8300	0.8310	0.8320	0.8329	0.8339	0.8348	0.8358	0.8368	0.8377
	cos	0.5592	0.5577	0.5563	0.5548	0.5534	0.5519	0.5505	0.5490	0.5476	0.5461
	tan	1.4826	1.4882	1.4938	1.4994	1.5051	1.5108	1.5166	1.5224	1.5282	1.5340
57	sin	0.8387	0.8396	0.8406	0.8415	0.8425	0.8434	0.8443	0.8453	0.8462	0.8471
	cos	0.5446	0.5432	0.5417	0.5402	0.5388	0.5373	0.5358	0.5344	0.5329	0.5314
	tan	1.5399	1.5458	1.5517	1.5577	1.5637	1.5697	1.5757	1.5818	1.5880	1.5941
58	sin	0.8480	0.8490	0.8499	0.8508	0.8517	0.8526	0.8536	0.8545	0.8554	0.8563
	cos	0.5299	0.5284	0.5270	0.5255	0.5240	0.5225	0.5210	0.5195	0.5180	0.5165
	tan	1.6003	1.6066	1.6128	1.6191	1.6255	1.6319	1.6383	1.6447	1.6512	1.6577
59	sin	0.8572	0.8581	0.8590	0.8599	0.8607	0.8616	0.8625	0.8634	0.8643	0.8652
	cos	0.5150	0.5135	0.5120	0.5105	0.5090	0.5075	0.5060	0.5045	0.5030	0.5015
	tan	1.6643	1.6709	1.6775	1.6842	1.6909	1.6977	1.7045	1.7113	1.7182	1.7251
Degs.	**Function**	**0′**	**6′**	**12′**	**18′**	**24′**	**30′**	**36′**	**42′**	**48′**	**54′**

NATURAL SINES, COSINES, AND TANGENTS

60°–74.9°

Degs.	Function	0.0°	0.1°	0.2°	0.3°	0.4°	0.5°	0.6°	0.7°	0.8°	0.9°
60	sin	0.8660	0.8669	0.8678	0.8686	0.8695	0.8704	0.8712	0.8721	0.8729	0.8738
	cos	0.5000	0.4985	0.4970	0.4955	0.4939	0.4924	0.4909	0.4894	0.4879	0.4863
	tan	1.7321	1.7391	1.7461	1.7532	1.7603	1.7675	1.7747	1.7820	1.7893	1.7966
61	sin	0.8746	0.8755	0.8763	0.8771	0.8780	0.8788	0.8796	0.8805	0.8813	0.8821
	cos	0.4848	0.4833	0.4818	0.4802	0.4787	0.4772	0.4756	0.4741	0.4726	0.4710
	tan	1.8040	1.8115	1.8190	1.8265	1.8341	1.8418	1.8495	1.8572	1.8650	1.8728
62	sin	0.8829	0.8838	0.8846	0.8854	0.8862	0.8870	0.8878	0.8886	0.8894	0.8902
	cos	0.4695	0.4679	0.4664	0.4648	0.4633	0.4617	0.4602	0.4586	0.4571	0.4555
	tan	1.8807	1.8887	1.8967	1.9047	1.9128	1.9210	1.9292	1.9375	1.9458	1.9542
63	sin	0.8910	0.8918	0.8926	0.8934	0.8942	0.8949	0.8957	0.8965	0.8973	0.8980
	cos	0.4540	0.4524	0.4509	0.4493	0.4478	0.4462	0.4446	0.4431	0.4415	0.4399
	tan	1.9626	1.9711	1.9797	1.9883	1.9970	2.0057	2.0145	2.0233	2.0323	2.0413
64	sin	0.8988	0.8996	0.9003	0.9011	0.9018	0.9026	0.9033	0.9041	0.9048	0.9056
	cos	0.4384	0.4368	0.4352	0.4337	0.4321	0.4305	0.4289	0.4274	0.4258	0.4242
	tan	2.0503	2.0594	2.0686	2.0778	2.0872	2.0965	2.1060	2.1155	2.1251	2.1348
65	sin	0.9063	0.9070	0.9078	0.9085	0.9092	0.9100	0.9107	0.9114	0.9121	0.9128
	cos	0.4226	0.4210	0.4195	0.4179	0.4163	0.4147	0.4131	0.4115	0.4099	0.4083
	tan	2.1445	2.1543	2.1642	2.1742	2.1842	2.1943	2.2045	2.2148	2.2251	2.2355
66	sin	0.9135	0.9143	0.9150	0.9157	0.9164	0.9171	0.9178	0.9184	0.9191	0.9198
	cos	0.4067	0.4051	0.4035	0.4019	0.4003	0.3987	0.3971	0.3955	0.3939	0.3923
	tan	2.2460	2.2566	2.2673	2.2781	2.2889	2.2998	2.3109	2.3220	2.3332	2.3445
67	sin	0.9205	0.9212	0.9219	0.9225	0.9232	0.9239	0.9245	0.9252	0.9259	0.9265
	cos	0.3907	0.3891	0.3875	0.3859	0.3843	0.3827	0.3811	0.3795	0.3778	0.3762
	tan	2.3559	2.3673	2.3789	2.3906	2.4023	2.4142	2.4262	2.4383	2.4504	2.4627
68	sin	0.9272	0.9278	0.9285	0.9291	0.9298	0.9304	0.9311	0.9317	0.9323	0.9330
	cos	0.3746	0.3730	0.3714	0.3697	0.3681	0.3665	0.3649	0.3633	0.3616	0.3600
	tan	2.4751	2.4876	2.5002	2.5129	2.5257	2.5386	2.5517	2.5649	2.5782	2.5916
69	sin	0.9336	0.9342	0.9348	0.9354	0.9361	0.9367	0.9373	0.9379	0.9385	0.9391
	cos	0.3584	0.3567	0.3551	0.3535	0.3518	0.3502	0.3486	0.3469	0.3453	0.3437
	tan	2.6051	2.6187	2.6325	2.6464	2.6605	2.6746	2.6889	2.7034	2.7179	2.7326
70	sin	0.9397	0.9403	0.9409	0.9415	0.9421	0.9426	0.9432	0.9438	0.9444	0.9449
	cos	0.3420	0.3404	0.3387	0.3371	0.3355	0.3338	0.3322	0.3305	0.3289	0.3272
	tan	2.7475	2.7625	2.7776	2.7929	2.8083	2.8239	2.8397	2.8556	2.8716	2.8878
71	sin	0.9455	0.9461	0.9466	0.9472	0.9478	0.9483	0.9489	0.9494	0.9500	0.9505
	cos	0.3256	0.3239	0.3223	0.3206	0.3190	0.3173	0.3156	0.3140	0.3123	0.3107
	tan	2.9042	2.9208	2.9375	2.9544	2.9714	2.9887	3.0061	3.0237	3.0415	3.0595
72	sin	0.9511	0.9516	0.9521	0.9527	0.9532	0.9537	0.9542	0.9548	0.9553	0.9558
	cos	0.3090	0.3074	0.3057	0.3040	0.3024	0.3007	0.2990	0.2974	0.2957	0.2940
	tan	3.0777	3.0961	3.1146	3.1334	3.1524	3.1716	3.1910	3.2106	3.2305	3.2506
73	sin	0.9563	0.9568	0.9573	0.9578	0.9583	0.9588	0.9593	0.9598	0.9603	0.9608
	cos	0.2924	0.2907	0.2890	0.2874	0.2857	0.2840	0.2823	0.2807	0.2790	0.2773
	tan	3.2709	3.2914	3.3122	3.3332	3.3544	3.3759	3.3977	3.4197	3.4420	3.4646
74	sin	0.9613	0.9617	0.9622	0.9627	0.9632	0.9636	0.9641	0.9646	0.9650	0.9655
	cos	0.2756	0.2740	0.2723	0.2706	0.2689	0.2672	0.2656	0.2639	0.2622	0.2605
	tan	3.4874	3.5105	3.5339	3.5576	3.5816	3.6059	3.6305	3.6554	3.6806	3.7062
Degs.	Function	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′

Degs.	Function	0.0°	0.1°	0.2°	0.3°	0.4°	0.5°	0.6°	0.7°	0.8°	0.9°
75	sin	0.9659	0.9664	0.9668	0.9673	0.9677	0.9681	0.9686	0.9690	0.9694	0.9699
	cos	0.2588	0.2571	0.2554	0.2538	0.2521	0.2504	0.2487	0.2470	0.2453	0.2436
	tan	3.7321	3.7583	3.7848	3.8118	3.8391	3.8667	3.8947	3.9232	3.9520	3.9812
76	sin	0.9703	0.9707	0.9711	0.9715	0.9720	0.9724	0.9728	0.9732	0.9736	0.9740
	cos	0.2419	0.2402	0.2385	0.2368	0.2351	0.2334	0.2317	0.2300	0.2284	0.2267
	tan	4.0108	4.0408	4.0713	4.1022	4.1335	4.1653	4.1976	4.2303	4.2635	4.2972
77	sin	0.9744	0.9748	0.9751	0.9755	0.9759	0.9763	0.9767	0.9770	0.9774	0.9778
	cos	0.2250	0.2232	0.2215	0.2198	0.2181	0.2164	0.2147	0.2130	0.2113	0.2096
	tan	4.3315	4.3662	4.4015	4.4374	4.4737	4.5107	4.5483	4.5864	4.6252	4.6646
78	sin	0.9781	0.9785	0.9789	0.9792	0.9796	0.9799	0.9803	0.9806	0.9810	0.9813
	cos	0.2079	0.2062	0.2045	0.2028	0.2011	0.1994	0.1977	0.1959	0.1942	0.1925
	tan	4.7046	4.7453	4.7867	4.8288	4.8716	4.9152	4.9594	5.0045	5.0504	5.0970
79	sin	0.9816	0.9820	0.9823	0.9826	0.9829	0.9833	0.9836	0.9839	0.9842	0.9845
	cos	0.1908	0.1891	0.1874	0.1857	0.1840	0.1822	0.1805	0.1788	0.1771	0.1754
	tan	5.1446	5.1929	5.2422	5.2924	5.3435	5.3955	5.4486	5.5026	5.5578	5.6140
80	sin	0.9848	0.9851	0.9854	0.9857	0.9860	0.9863	0.9866	0.9869	0.9871	0.9874
	cos	0.1736	0.1719	0.1702	0.1685	0.1668	0.1650	0.1633	0.1616	0.1599	0.1582
	tan	5.6713	5.7297	5.7894	5.8502	5.9124	5.9758	6.0405	6.1066	6.1742	6.2432
81	sin	0.9877	0.9880	0.9882	0.9885	0.9888	0.9890	0.9893	0.9895	0.9898	0.9900
	cos	0.1564	0.1547	0.1530	0.1513	0.1495	0.1478	0.1461	0.1444	0.1426	0.1409
	tan	6.3138	6.3859	6.4596	6.5350	6.6122	6.6912	6.7720	6.8548	6.9395	7.0264
82	sin	0.9903	0.9905	0.9907	0.9910	0.9912	0.9914	0.9917	0.9919	0.9921	0.9923
	cos	0.1392	0.1374	0.1357	0.1340	0.1323	0.1305	0.1288	0.1271	0.1253	0.1236
	tan	7.1154	7.2066	7.3002	7.3962	7.4947	7.5958	7.6996	7.8062	7.9158	8.0285
83	sin	0.9925	0.9928	0.9930	0.9932	0.9934	0.9936	0.9938	0.9940	0.9942	0.9943
	cos	0.1219	0.1201	0.1184	0.1167	0.1149	0.1132	0.1115	0.1097	0.1080	0.1063
	tan	8.1443	8.2636	8.3863	8.5126	8.6427	8.7769	8.9152	9.0579	9.2052	9.3572
84	sin	0.9945	0.9947	0.9949	0.9951	0.9952	0.9954	0.9956	0.9957	0.9959	0.9960
	cos	0.1045	0.1028	0.1011	0.0993	0.0976	0.0958	0.0941	0.0924	0.0906	0.0889
	tan	9.5144	9.6768	9.8448	10.02	10.20	10.39	10.58	10.78	10.99	11.20
85	sin	0.9962	0.9963	0.9965	0.9966	0.9968	0.9969	0.9971	0.9972	0.9973	0.9974
	cos	0.0872	0.0854	0.0837	0.0819	0.0802	0.0785	0.0767	0.0750	0.0732	0.0715
	tan	11.43	11.66	11.91	12.16	12.43	12.71	13.00	13.30	13.62	13.95
86	sin	0.9976	0.9977	0.9978	0.9979	0.9980	0.9981	0.9982	0.9983	0.9984	0.9985
	cos	0.0698	0.0680	0.0663	0.0645	0.0628	0.0610	0.0593	0.0576	0.0558	0.0541
	tan	14.30	14.67	15.06	15.46	15.89	16.35	16.83	17.34	17.89	18.46
87	sin	0.9986	0.9987	0.9988	0.9989	0.9990	0.9990	0.9991	0.9992	0.9993	0.9993
	cos	0.0523	0.0506	0.0488	0.0471	0.0454	0.0436	0.0419	0.0401	0.0384	0.0366
	tan	19.08	19.74	20.45	21.20	22.02	22.90	23.86	24.90	26.03	27.27
88	sin	0.9994	0.9995	0.9995	0.9996	0.9996	0.9997	0.9997	0.9997	0.9998	0.9998
	cos	0.0349	0.0332	0.0314	0.0297	0.0279	0.0262	0.0244	0.0227	0.0209	0.0192
	tan	28.64	30.14	31.82	33.69	35.80	38.19	40.92	44.07	47.74	52.08
89	sin	0.9998	0.9999	0.9999	0.9999	0.9999	1.000	1.000	1.000	1.000	1.000
	cos	0.0175	0.0157	0.0140	0.0122	0.0105	0.0087	0.0070	0.0052	0.0035	0.0017
	tan	57.29	63.66	71.62	81.85	95.49	114.6	143.2	191.0	286.5	573.0
Degs.	Function	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′

APPENDIX VII

COMMON LOGARITHMS OF NUMBERS

No.	0	1	2	3	4	5	6	7	8	9
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222
42	6232	6243	6253	6263	6274	6284	6294	6304	6314	6325
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396
No.	0	1	2	3	4	5	6	7	8	9

COMMON LOGARITHMS OF NUMBERS

No.	0	1	2	3	4	5	6	7	8	9
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7627
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122
65	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289
85	9294	9299	9304	9309	9315	9320	9325	9330	9335	9340
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952
99	9956	9961	9965	9969	9974	9978	9983	9987	9991	9996
No.	0	1	2	3	4	5	6	7	8	9

APPENDIX VIII

ANSWERS TO QUIZZES

CHAPTER 1

DUTIES AND RESPONSIBILITIES

1. a.
2. b.
3. a.
4. c.
5. a.
6. b.
7. b.
8. c.
9. c.
10. b.
11. a.
12. b.
13. c.
14. c.
15. a.
16. d.

CHAPTER 2

R–C AND R–L NETWORKS

1. a.
2. c.
3. d.
4. a.
5. b.
6. d.
7. Yes
8. d.
9. a.
10. b.
11. No difference
12. Yes
13. a.
14. d.
15. c.
16. b.
17. b.
18. b.
19. a.
20. b.
21. a.

CHAPTER 3

SPECIAL CIRCUITS

1. b.
2. b.
3. c.
4. a.
5. b.
6. d.
7. d.
8. a.
9. c.
10. d.
11. b.
12. d.
13. d.
14. c.
15. d.
16. a.
17. b.
18. b.
19. d.
20. d.
21. b.
22. c.
23. a.
24. d.

CHAPTER 4

SPECIAL CIRCUITS—CONTINUED

1. a.
2. c.
3. c.
4. c.
5. b.
6. d.
7. a.
8. d.
9. c.
10. d.
11. b.
12. c.
13. c.
14. b.
15. d.
16. c.
17. b.
18. a.
19. b.
20. a.
21. b.
22. d.

CHAPTER 5

WAVEGUIDES AND CAVITY RESONATORS

1. a.
2. b.
3. c.
4. b.
5. b.
6. d.
7. b.
8. b.
9. a.
10. c.
11. d.
12. a.
13. d.
14. a.
15. b.
16. b.
17. c.
18. b.

CHAPTER 6

UHF AND MICROWAVE TECHNIQUES

1. a.
2. a.
3. d.
4. c.
5. a.
6. c.
7. a.
8. d.
9. b.
10. b.
11. c.
12. a.
13. a.
14. c.
15. a.
16. c.
17. b.

CHAPTER 7

UHF AND MICROWAVE ANTENNAS

1. a.
2. c.
3. b.
4. d.
5. b.
6. a.
7. a.
8. c.
9. d.
10. b.
11. a.
12. c.
13. b.
14. c.
15. a.
16. b.
17. c.
18. a.
19. b.
20. b.
21. c.
22. d.
23. c.
24. a.

CHAPTER 8

SERVOMECHANISMS

1. d.
2. c.
3. c.
4. b.
5. a.
6. a.
7. d.
8. b.
9. c.
10. a.
11. b.
12. c.
13. b.
14. d.
15. d.
16. a,
17. a.
18. c.
19. d.
20. b.
21. a.
22. b.
23. d.
24. a.

CHAPTER 9

MAGNETIC AMPLIFIERS

1. c.
2. a.
3. a.
4. d.
5. b.
6. d.
7. c.
8. d.
9. a.
10. a.
11. b.
12. a.
13. d.
14. c.
15. a.
16. c.
17. a.
18. d.
19. a.
20. b.
21. c.

CHAPTER 10

INTRODUCTION TO TRANSISTORS

1. c.
2. b.
3. c.
4. a.
5. b.
6. d.
7. c.
8. b.
9. d.
10. a.
11. c.
12. b.
13. a.
14. c.
15. b.
16. d.
17. a.
18. b.
19. c.
20. c.
21. b.
22. c.
23. a.
24. a.
25. d.

CHAPTER 11

SPECIAL MAINTENANCE TECHNIQUES

1. c.
2. b.
3. d.
4. a.
5. c.
6. d.
7. d.
8. c.
9. a.
10. c.
11. b.
12. a.
13. b.
14. c.
15. d.
16. a.
17. a.
18. c.
19. b.
20. d.
21. d.
22. d.
23. d.
24. b.

CHAPTER 12

AVIATION ELECTRONICS SUPPLY

1. c.
2. b.
3. c.
4. a.
5. b.
6. b.
7. a.
8. c.
9. d.
10. d.
11. a.
12. d.

APPENDIX IX

QUALIFICATIONS FOR ADVANCEMENT IN RATING

AVIATION ELECTRONICS TECHNICIANS (AT)

RATING CODE NO. 6300

General Service Rating

Scope

Aviation Electronics Technicians test, maintain, and repair aviation electronic equipment; inspect, clean, lubricate, and make operational tests and adjustments of communication, navigation, search, identification, electronic countermeasures, detection, and relay equipment; remove and reinstall components, assemblies, and subassemblies; calibrate, repair, and make performance measurements; make detailed mechanical, electrical, and electronic casualty analysis; aline communication, navigation, search, identification, display and relay equipments; make authorized repairs and adjustments to associated test equipment.

Emergency Service Ratings

AVIATION ELECTRONICS TECHNICIANS R (Radar), Rating Code No. 6304--ATR

Test, maintain, and repair aviation electronic equipment including radar, radar altimeters, radar display, radar relay and identification equipment; inspect, clean, lubricate, and make operational tests and adjustments of search, identification, radar detection, and relay equipments; remove and install/reinstall components, assemblies, subassemblies, and detailed parts; calibrate, repair, and make performance measurements; make detailed mechanical, electrical, and electronic casualty analysis; aline radar search, display and relay equipments and identification equipment; make authorized repairs and adjustments to associated test equipment.

AVIATION ELECTRONICS TECHNICIANS S (ASW), Rating Code No. 6305 -- ATS

Test, aline, maintain, and repair aviation electronic anti-submarine detection and reconnaissance equipment other than radar; inspect, clean, lubricate, and make operational tests and adjustments of electronic countermeasures (ECM), magnetic airborne detection (MAD), sonobuoys, and related equipments and airborne sonar and related equipments; remove, install/reinstall components, assemblies, subassemblies, and detailed parts; calibrate, repair, and make performance measurements; make detailed mechanical, electrical, and electronic casualty analysis; make authorized repairs and adjustments to associated test equipment.

AVIATION ELECTRONICS TECHNICIANS N (Communications and Navigation Equipment), Rating Code No. 6306 ------ ATN

Test, maintain, and repair aviation electronic communication and navigation equipment; inspect, clean, lubricate, and make operational tests of communications and navigation equipments; remove and install/reinstall components, assemblies, subassemblies, and detailed parts; calibrate, repair, and make performance measurements; make detailed mechanical, electrical, and electronic casualty analysis; make authorized repairs and adjustments to associated test equipment.

Navy Enlisted Classification Codes

For specific Navy enlisted classification codes included within this rating, see *Manual of Navy Enlisted Classifications,* NavPers 15105 (Revised), Codes AT–6600 through AT–6699.

Qualifications for advancement in rating		*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
100	PRACTICAL FACTORS				
101	OPERATIONAL				
	1. Demonstrate under simulated conditions the rescue of a person in contact with an energized electrical circuit, resuscitation of a person unconscious from electrical shock, and treatment for electrical burns	3	3	3	3
	2. Energize, secure, set operating controls, make operator's adjustments, and read and interpret dials, built-in meters, and cathode ray tube presentations on:				
	*a. Aviation radar equipment	3	3	-----	-----
	*b. Aviation ASW equipment	3	-----	3	-----
	*c. Aviation communications and navigation equipment	3	-----	-----	3
102	MAINTENANCE AND/OR REPAIR				
	1. Select and use hand tools and small portable power tools provided for maintenance and repair of:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	2. Inspect and clean commutators and slip ring assemblies; inspect and replace brushes	3	3	3	3
	3. Lubricate and clean the following in accordance with aviation electronic technical and maintenance publications:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3

*Where terms "aviation radar equipment", "aviation ASW equipment", and "aviation communications and navigation equipment" are used in the qualifications items, these equipments are defined in terms of the scopes of emergency service ratings.

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	ATR	ATS	ATN
102	MAINTENANCE AND/OR REPAIR—Cont.				
	4. Operate the following test equipment in performance of line maintenance:				
	a. Ohmmeter	3	3	3	3
	b. Ammeter	3	3	3	3
	c. Voltmeter	3	3	3	3
	d. Megger	3	3	3	3
	e. Tube tester	3	3	3	3
	f. Frequency meter	3	3	3	3
	g. Multimeter	3	3	3	3
	h. Oscilloscope	3	3	3	3
	i. RF signal generator	3	3	3	3
	j. AF signal generator	3	3	3	3
	k. Sonobuoy test set	-----	-----	3	-----
	l. Magnetic detection test set	-----	-----	3	-----
	m. Output meter	-----	-----	-----	3
	5. Make tests on the following for short circuits, grounds, and continuity of interconnecting cables between components, and report results:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	6. Localize equipment casualties in aircraft to removable component in:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	7. Make electrical connections and splices, including soldered joints	3	3	3	3
	8. Remove, reinstall, inspect and test for correct installation in aircraft, components of:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	9. Test and replace electron tubes, fuses, wiring, lamps, and connectors (excluding potting process)	3	3	3	3

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
102	MAINTENANCE AND/OR REPAIR—Cont.				
	10. Draw and interpret schematic diagrams of electrical circuits; read and interpret wiring diagrams of electrical circuits found in aircraft and equipment maintenance publications where applicable to maintenance and repair of:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	11. Locate and identify components of the following electronic equipment in aircraft by reference to block diagrams and mechanical drawings:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	12. Operate the following test equipment in performance of shop maintenance:				
	a. Vacuum tube voltmeter	2	2	2	2
	b. Spectrum analyzer	2	2	-----	-----
	c. Signal generators	2	2	2	2
	d. Frequency meters	2	2	2	2
	e. RF wattmeter	2	2	2	2
	f. Echo box	2	2	-----	-----
	13. Test the following aviation electronic equipment for short circuits, grounds, and continuity in electronic circuits; measure electrical quantities, such as voltage, current, power, frequency, and phase angle, and compare with established values; trace signals, using an oscilloscope; determine wave forms and compare with established or required forms:				
	a. Aviation radar equipment	2	2	-----	-----
	b. Aviation ASW equipment	2	-----	2	-----
	c. Aviation communications and navigation equipment	2	-----	-----	2

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
102	MAINTENANCE AND/OR REPAIR—Cont.				
	14. Localize equipment casualties to detail parts, such as: Resistors, plug-in units, potentiometers, synchros, and tubes within a component, assembly, and subassembly in:				
	a. Aviation radar equipment	2	2	-----	-----
	b. Aviation ASW equipment	2	-----	2	-----
	c. Aviation communications and navigation equipment	2	-----	-----	2
	15. Effect alterations or modifications to aviation electronic material in accordance with Aircraft Service Changes and Electronic Material Changes	2	2	2	2
	16. Repair and adjust aviation electronic equipment circuitry by replacement or repair of assemblies, subassemblies, or detailed parts	2	2	2	2
	17. Read and interpret electronic wiring and schematic diagrams, waveform charts, voltage and resistance charts in aviation electronic technical and maintenance publications	2	2	2	2
	18. Evaluate test equipment for correct operations: Make authorized repairs and calibrations	1	1	1	1
	19. Diagnose and localize casualties within the following systems:				
	a. Aviation radar systems	1	1	-----	-----
	b. Aviation ASW systems	1	-----	1	-----
	c. Aviation communications and navigation systems	1	-----	-----	1
	20. Adjust and aline parts or circuitry in components, assemblies, or subassemblies to conform to allowable limits	1	1	1	1
	21. Accomplish quantitative performance checks on systems and compare with prescribed standards	1	1	1	1

	Qualifications for advancement in rating	APPLICABLE RATES			
		AT	ATR	ATS	ATN
102	MAINTENANCE AND/OR REPAIR—Cont.				
	22. Evaluate electrical and electronic tests made on aircraft, aviation electronic equipment, components, assemblies, and subassemblies; reduce data and make corrections; compare values against prescribed performance criteria	C	C	C	C
	23. Evaluate repaired, overhauled, or newly installed aviation electronic radar equipment or components thereof for proper performance	C	C	-----	-----
	24. Evaluate repaired, overhauled, or newly installed aviation electronic ASW equipment or components thereof for proper performance	C	-----	C	-----
	25. Evaluate repaired, overhauled, or newly installed aviation electronic communications and navigation equipment	C	-----	-----	C
	26. Screen for feasibility of local repairs: Equipment, components, assemblies, subassemblies, or parts recommended for exchange	C	C	C	C
103	ADMINISTRATIVE AND/OR CLERICAL				
	1. Record test data and work accomplished in required work logs; complete check lists	3	3	3	3
	2. Prepare reports of equipment failure; prepare shop requisitions for spare parts, tools, and material needed to repair:				
	a. Aviation radar equipment	2	2	-----	-----
	b. Aviation ASW equipment	2	-----	2	-----
	c. Aviation communications and navigation equipment	2	-----	-----	2

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
103	ADMINISTRATIVE AND/OR CLERICAL				
	3. Evaluate shop and equipment logs, completed equipment check lists, reports of equipment failure; requisitions for spare parts, tools, and material in repairing:				
	a. Aviation radar equipment	1	1	-----	-----
	b. Aviation ASW equipment	1	-----	1	-----
	c. Aviation communications and navigation equipment	1	-----	-----	1
	4. Take, record, and report inventory of spare parts, tools, test equipment, and materials in shop and shop storeroom	1	1	1	1
	5. Train and supervise personnel in the maintenance and repair of:				
	a. Aviation radar equipment	1	1	-----	-----
	b. Aviation ASW equipment	1	-----	1	-----
	c. Aviation communications and navigation equipment	1	-----	-----	1
	6. Organize and administer a shop for maintenance and repair of:				
	a. Aviation radar equipment	C	C	-----	-----
	b. Aviation ASW equipment	C	-----	C	-----
	c. Aviation communications and navigation equipment	C	-----	-----	C
	7. Prepare reports covering shop work accomplished, materials used, and the status and condition of aviation electronic equipment and test equipment	C	C	C	C
	8. Evaluate inventories in terms of shop requirements and prepare requisitions to replenish stock or to obtain new items	C	C	C	C
200	EXAMINATION SUBJECTS				
201	OPERATIONAL				
	1. Effects of electrical shock, method of resuscitation of a person unconscious from electrical shock, and treatment for electrical burns	3	3	3	3

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	ATR	ATS	ATN
201	OPERATIONAL—Cont.				
	2. Safety precautions to be observed when working on the line and in a hangar or shop, as set forth in chapter 3, *U.S. Navy Safety Precautions* (OpNav 34P1)	3	3	3	3
	3. Electrical and electronic safety precautions applicable to aviation personnel as set forth in chapter 18, *U.S. Navy Safety Precautions* (OpNav 34P1)	3	3	3	3
202	MAINTENANCE AND/OR REPAIR				
	1. Types and purpose of hand tools and small portable power tools provided for use in the maintenance and repair of:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	2. Function of the following test equipment:				
	a. Ohmmeter	3	3	3	3
	b. Ammeter	3	3	3	3
	c. Voltmeter	3	3	3	3
	d. Megger	3	3	3	3
	e. Multimeter	3	3	3	3
	f. Tube tester	3	3	3	3
	g. Echo box	3	3	-----	-----
	h. Frequency meters (portable line maintenance type)	3	3	3	3
	i. Oscilloscope	3	3	3	3
	j. RF signal generator	3	3	3	3
	k. AF signal generator	3	3	3	3
	l. Crystal checker	3	3	3	-----
	m. Sonobuoy test set	-----	-----	3	-----
	n. Magnetic detection test set	-----	-----	3	-----
	o. Output meter	-----	-----	-----	3

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	ATR	ATS	ATN
202	MAINTENANCE AND/OR REPAIR—Cont.				
	3. Operating principles of voltmeter, ammeter, and ohmmeter (but not vacuum tube voltmeter), including:				
	a. D'Arsonval meter movement	3	3	3	3
	b. Shunts and multipliers	3	3	3	3
	c. Use of thermocouples and rectifiers in a-c meters	3	3	3	3
	4. Methods and equipment used in electrical tests for continuity, grounds, and short circuits	3	3	3	3
	5. Soldering materials and methods of soldering used in maintenance and repair of:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	6. Types of information shown and meaning of electrical and electronic symbols used in schematic and block diagrams of:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	7. Units of electrical measurement, including:				
	a. Volt	3	3	3	3
	b. Ampere	3	3	3	3
	c. Ohm	3	3	3	3
	d. Watt	3	3	3	3
	e. Volt-ampere	3	3	3	3
	f. Henry	3	3	3	3
	g. Farad	3	3	3	3
	h. Cycle	3	3	3	3
	i. Ampere-turn	3	3	3	3
	j. Prefixing of mega, kilo, milli, and micro	3	3	3	3
	k. Gamma	-----	-----	3	-----
	l. Oersted	-----	-----	3	-----
	m. Gauss	-----	-----	3	-----

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
202	MAINTENANCE AND/OR REPAIR—Cont.				
	8. Meaning of:				
	a. Conductors and insulators	3	3	3	3
	b. Field intensity	3	3	3	3
	c. Flux density	3	3	3	3
	d. Permeability	3	3	3	3
	e. Hysteresis and eddy currents	3	3	3	3
	f. Self and mutual inductance	3	3	3	3
	g. Electromagnetic induction	3	3	3	3
	h. Power factor	3	3	3	3
	i. Reactance	3	3	3	3
	j. Impedance	3	3	3	3
	k. Capacitance	3	3	3	3
	l. Frequency	3	3	3	3
	m. Phase	3	3	3	3
	n. Time constant	3	3	3	3
	9. System of assigning "AN" letter-number combinations as designation for electronic equipment	3	3	3	3
	10. RMA color coding systems for capacitors and resistors	3	3	3	3
	11. Operating principles of superheterodyne receivers	3	3	3	3
	12. Function of the following in electrical/electronic circuits:				
	a. Resistors	3	3	3	3
	b. Rheostats and potentiometers	3	3	3	3
	c. Solenoids	3	3	3	3
	d. Inductors	3	3	3	3
	e. Capacitors	3	3	3	3
	f. Fuses	3	3	3	3
	g. Switches	3	3	3	3
	h. Reactors	3	3	3	3
	i. Transformers	3	3	3	3
	j. Relays	3	3	3	3
	k. Synchros	-----	3	3	-----
	l. Servomechanisms	-----	3	3	-----
	13. Types of electron tubes used in electronic circuits and operating principles of diode, triode, tetrode, pentode, and mixer/converter tubes	3	3	3	3

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
202	MAINTENANCE AND/OR REPAIR—Cont.				
	14. Relationship of resistance; inductance, and capacitance in a-c circuits	3	3	3	3
	15. Relationship of current, voltage, and impedance in a-c circuits	3	3	3	3
	16. Relationship of the length and cross-sectional area to resistance of a conductor	3	3	3	3
	17. Relationship of resistance, temperature, and current in an electrical conductor	3	3	3	3
	18. Calculate current, voltage, and resistance in d.c. series and parallel circuits containing not more than four elements	3	3	3	3
	19. Component parts of motors, generators, and alternators; application of laws of magnetism to electrical rotating machinery	3	3	3	3
	20. Effects of meter sensitivity in circuit voltage measurement	3	3	3	3
	21. Function and operating principles of the following:				
	a. D-c power supplies (half-wave, full wave, and bridge-type rectifiers, voltage doublers, capacitor and choke input filters)	3	3	3	3
	b. D-c voltage regulators	3	3	3	3
	22. Function of the following shop type test equipment:				
	a. Vacuum tube voltmeter	2	2	2	2
	b. Spectrum analyzer	2	2	-----	-----
	c. Signal generators	2	2	2	2
	d. Frequency meters	2	2	2	2
	e. RF wattmeter	2	2	2	2
	f. MAD equipment analyzer	-----	-----	2	-----
	g. Sonar test set	-----	-----	2	-----
	h. Permeability shield cans	-----	---	2	---
	i. Magnet set	-----	-----	2	-----
	23. Circuit loading effects of test equipment	2	2	2	2

Qualifications for advancement in rating		*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
202	MAINTENANCE AND/OR REPAIR—Cont.				
	24. Function and operating principles of the following electronic circuits:				
	a. Audio amplifiers	**2**	**2**	**2**	**2**
	b. Video amplifiers	**2**	**2**	**2**	**2**
	c. RF amplifiers	**2**	**2**	**2**	**2**
	d. IF amplifiers	**2**	**2**	**2**	**2**
	e. Cathode followers	**2**	**2**	**2**	**2**
	f. Oscillators: (Hartley, crystal-controlled, Colpitts, TPTG, and electron-coupled)	**2**	**2**	**2**	**2**
	g. Detectors: (diode and crystal)	**2**	**2**	**2**	**2**
	h. Mixer/converter	**2**	**2**	**2**	**2**
	i. Frequency multipliers	**2**	**2**	**2**	**2**
	j. Modulators	**2**	**2**	**2**	**2**
	25. Relationship of current, voltage, and impedance in series and parallel resonant circuits at, above, and below resonance	**2**	**2**	**2**	**2**
	26. Relationship of reluctance, flux, and magnetomotive force in a-c and d-c magnetic circuits	**2**	**2**	**2**	**2**
	27. Calculate current, voltage, phase angle, impedance, and resonance in a-c series and parallel circuits containing not more than four elements	**2**	**2**	**2**	**2**
	28. Operating principles of test equipment authorized for repair and/or calibration of:				
	a. Aviation radar equipment	**1**	**1**	-----	-----
	b. Aviation ASW equipment	**1**	-----	**1**	-----
	c. Aviation communications and navigation equipment	**1**	-----	-----	**1**
	29. Methods and equipment used in adjustment and alinement of parts or circuitry in components, assemblies, or subassemblies, of the following, to conform to allowable limits:				
	a. Aviation radar equipment	**1**	**1**	-----	-----
	b. Aviation ASW equipment	**1**	-----	**1**	-----
	c. Aviation communications and navigation equipment	**1**	-----	-----	**1**

Qualifications for advancement in rating		*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
202	MAINTENANCE AND/OR REPAIR—Cont.				
	30. Operating principles of the following:				
	a. Tuned coupling circuits	1	1	1	1
	b. Impedance matching	1	1	1	1
	c. Phase shifters	1	1	1	1
	d. Differentiators	1	1	1	1
	e. Integrators	1	1	1	1
	f. Oscillators: (blocking and Wein-Bridge)	1	1	1	1
	g. Trigger circuits and multivibrators	1	1	1	1
	h. Coincidence circuits	1	1	1	1
	i. Limiters: (noise and pulse)	1	1	1	1
	j. Clippers	1	1	1	1
	k. Discriminators	1	1	1	1
	l. AGC and AFC circuits	1	1	1	1
	m. Squelch circuits	1	1	1	1
	n. Counting circuits	1	1	1	1
	o. Sawtooth generators	1	1	1	1
	p. Peakers	1	1	1	1
	q. Clampers	1	1	1	1
	r. Delay lines	1	1	1	1
	s. Magnetic amplifiers	1	1	1	1
	31. Operating principles of amplitude, frequency, phase, and pulse modulation	1	1	1	1
	32. Application and operating principles of:				
	a. Wire and coaxial transmission lines	1	1	1	1
	b. Wave guides	1	1	-----	-----
	c. T/R tubes	1	1	-----	-----
	d. AT/R tubes	1	1	-----	-----
	e. Klystrons	1	1	1	-----
	f. Magnetrons	1	1	-----	-----
	g. Crystal mixers	1	1	1	1
	h. Magnetic amplifiers	1	1	1	1
	33. Function and operating principles of gas-filled electron and cathode ray tubes	1	1	1	1

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
202	MAINTENANCE AND/OR REPAIR—Cont.				
	34. Methods of making RF power measurements on:				
	a. Aviation radar equipment	1	1	-----	-----
	b. Aviation ASW equipment	1	-----	1	-----
	c. Aviation communications and navigation equipment	1	-----	-----	1
	35. Methods of coupling: Transformer, impedance, capacitance, resistive, and direct	1	1	1	1
	36. Functions and applications of servo mechanisms and synchros as applied to aviation electronic equipment	1	1	1	1
	37. Method of making gain, phase, balancing, and zeroing adjustments to servo loops	1	1	1	1
	38. Electrical characteristics of Hertz, Marconi, cone, sleeve, plate, slot, and dipole antennas; operating principles of dish reflectors and associated RF feeds and of parasitic reflectors and directors	C	C	C	C
	39. Factors causing and methods of suppressing or eliminating radio noise interference	C	C	C	C
	40. Cause and/or effect of induction field, radiation field, sky wave, ground wave, ground reflected wave, ionospheric reflecting layers, skip distances, and maximum usable frequency	C	C	C	C
	41. Methods of measuring field strength of radio waves	C	C	C	C
	42. Effects of various combinations of beam widths, pulse length, pulse repetition frequency, and power output relative to performance and applications of radar equipment	C	C	-----	-----
	43. Effects of environmental conditions upon operation of electronic and electrical equipment, and special maintenance techniques involved in equipments to be operated at extremes of altitude, temperature, and humidity	C	C	C	C

	Qualifications for advancement in rating	*APPLICABLE RATES*			
		AT	**ATR**	**ATS**	**ATN**
203	ADMINISTRATIVE AND/OR CLERICAL				
	1. Types of forms and records used in maintenance and repair of:				
	a. Aviation radar equipment	3	3	-----	-----
	b. Aviation ASW equipment	3	-----	3	-----
	c. Aviation communications and navigation equipment	3	-----	-----	3
	2. Types of information found in aviation electronic technical and maintenance publications	3	3	3	3
	3. Types of entries made in electronics portion of Standard Aircraft Inventory Log; types of information recorded in reports of equipment failure	2	2	2	2
	4. Types of information found in Electronic Material Bulletins and Changes	2	2	2	2
	5. Purpose and type of information contained in Section "R" allowance list of aviation electronic material	2	2	2	2
	6. Procedures for obtaining replacement parts, tools, test equipment and materials, within the shop for:				
	a. Aviation radar equipment	1	1	-----	-----
	b. Aviation ASW equipment	1	-----	1	-----
	c. Aviation communications and navigation equipment	1	-----	-----	1
	7. Types of information found in Aviation Circular Letters, Aircraft Bulletins, and Aircraft Service Changes	1	1	1	1
	8. Application of allowance lists in determining spare parts, tools, and supplies authorized for own activity	1	1	1	1
	9. Procedures to be used for obtaining replacement parts, materials, authorized tools, and test equipment for the shop, and for returning defective equipment to supply system for processing to overhaul and repair facilities	C	C	C	C
	10. Procedures for survey of accountable material	C	C	C	C

Qualifications for advancement in rating		APPLICABLE RATES			
		AT	ATR	ATS	ATN
300	PATH OF ADVANCEMENT TO WARRANT OFFICER AND LIMITED DUTY OFFICER Aviation Electronics Technicians advance to Warrant Aviation Electronics Technician and/or to Limited Duty Officer, Aviation Electronics.				

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